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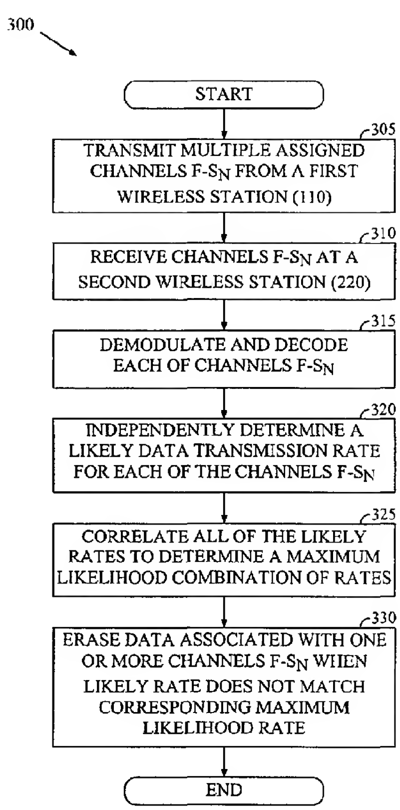
# (54) Title: METHOD AND APPARATUS FOR IS-95B REVERSE LINK SUPPLEMENTAL CODE CHANNEL FRAME VALIDATION AND FUNDAMENTAL CODE CHANNEL RATE DECISION IMPROVEMENT




**(57) Abstract:** The present invention provides a method and apparatus for maximizing throughput of a data call in a wireless communication system in which data is transmitted from a wireless station, such as a mobile station, on multiple assigned channels in accordance with a known transmission standard, such as IS-95B. The multiple assigned channels include a fundamental channel and at least one supplemental channel. Data is formatted into variable rate data frames and transmitted on the fundamental channel and the supplemental channel. A wireless receiver, such as a base station, receives the multiple assigned channels. The wireless receiver demodulates and decodes data frames associated with each of the multiple assigned channels. The wireless receiver determines a likely initial data rate for each demodulated and decode data frame. The wireless receiver correlates all of the likely data rates, by comparison to one another and to a relevant transmission protocol standard, to determine a maximum likelihood combination of data rates. The maximum likelihood combination of data rates includes a maximum likelihood data rate corresponding to each likely data rate. Decoded data frames are invalidated and erased when the likely data frame rates do not match corresponding maximum likelihood data rates.

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# METHOD AND APPARATUS FOR IS-95B REVERSE LINK SUPPLEMENTAL CODE CHANNEL FRAME VALIDATION AND FUNDAMENTAL CODE CHANNEL RATE DECISION IMPROVEMENT

## BACKGROUND OF THE INVENTION

### Field of the Invention

The present invention relates generally to wireless communication systems, and more particularly, to such a system for maximizing the useful data transmission throughput in a data call in which data is transmitted between wireless stations on multiple assigned channels.

### Related Art

A wireless communication system can be used to transmit synchronous and asynchronous packet data between a wireless transmitter and a wireless receiver. For example, the wireless communication system can operate in accordance with a High Speed Packet Data (HSPD) feature of the "TIA/EIA/IS-95B Mobile Station-Base Station Compatibility Standard for Dual-mode Wideband Spread Spectrum Cellular Systems" (hereinafter referred to as IS-95B) to achieve a packet data transmission bandwidth of up to 115 kilobits-per-second (kbps). Under IS-95B, a mobile station can transmit data to a base station receiver on an IS-95B reverse-link traffic channel including a fundamental code channel (FCCH) and up to seven additional Supplemental Code Channels (SCCHs). The FCCH is a variable rate channel capable of operating at data transmission rates including a full rate, a half rate, a quarter rate, and an eighth rate. On the other hand, the SCCH operates only at a full rate when data is to be transmitted, and at a zero rate when no data is available.

Packet data transmitted on the FCCH and SCCHs is partitioned into 20 millisecond (ms) variable rate data frames. Although the data rate can change rapidly, for example, on a frame by frame basis, rate information is typically not included in each transmitted data frame for at least two reasons. First, including rate information in each data frame wastes data bandwidth, and second, corruption of such transmitted rate information would adversely affect the entire frame. Since rate information is not included in each transmitted data frame, the receiver must determine from each received data frame (without the aid of embedded rate information) the rate at which the frame was transmitted, to thereby enable the receiver to properly process the data in the data frame. Known methods of determining data frame rates exist for voice only traffic. However, such methods are insufficiently accurate and thus unsuitable for packet data traffic.

Therefore, there is a need in a variable rate communication system to accurately determine a transmitted data rate for packet data traffic at a receiver without embedding rate information into the transmitted data.

In the above described communication system, the mobile station sends signaling requests for SCCH assignment and de-assignment to the base station based on the amount of data the mobile station needs to transmit. In response, the base station dynamically allocates and de-allocates SCCHs via signaling messages. Assigning and de-assigning SCCHs via such signaling can be a relatively slow mechanism and thus wastes valuable data transmission bandwidth. For example, assigning or de-assigning an SCCH can take up to a half-second.

To reduce assignments and de-assignments and associated delays, a mobile station can operate in a discontinuous transmission (DTX) mode while a SCCH is assigned to the mobile station. The DTX mode permits the mobile station to stop transmitting on the assigned SCCH while data is unavailable. This is referred to as the DTX "black-out" period. The DTX mode also permits the mobile station to resume transmitting as soon as data becomes available, thus avoiding delays associated with assigning and de-assigning the SCCH.

Transmitted data frames typically do not include DTX "on/off" information for similar reasons as mentioned above with regard to rate information. Since the receiver of the assigned SCCH receives no explicit indicator regarding the black-out periods, the receiver continuously demodulates and decodes the SCCH as long as the SCCH is assigned, even during the black-out period when no data is being transmitted, that is, when the demodulated and decoded data is invalid.

Therefore, it is desirable at a receiver in a communication system to discriminate between data transmission periods and black-outs so as to reduce a likelihood that invalid data is declared to be valid at the receiver.

In accordance with IS-95B, each transmitted SCCH data frame includes a 12 bit Cyclic Redundancy Code (CRC) for checking the validity of the data in the data frame at the receiver. Additional observable metrics, such as a Yamamoto measure, a symbol error rate, a frame energy, and so on, can be used to further improve on the CRC check. There is a finite probability ($2^{-12} = 2.4 \times 10^{-4}$) that demodulated random data associated with the black-out period, or noise corrupting a received data frame, will cause an erroneous match of the 12 bit CRC. In the case of a black-out period, a non-existent SCCH data frame or "random frame" corresponding to the erroneous CRC match, erroneously labels the invalid random frame as a valid data frame.

As is known, the transmitter and receiver typically implement complementary or parallel, layered, communication protocol layers including a physical protocol layer and an overlaying Radio Link Protocol (RLP) layer. One known RLP layer useable in wireless data communication stations is the IS-707 Radio Link Protocol. The physical layer sends (and receives) supposedly valid data frames (for example, data frames passing the CRC check as mentioned above) to (and from) the RLP. The RLP at the receiver tracks RLP frame sequence numbers embedded in the data frames for purposes of errored frames re-transmission and control.

During black out-periods, it has been observed that passing random frames as valid data frames to the RLP causes the RLP to initiate error control

processes. This can occur on either the FCCH or SCCHs. For example, the RLP will reset and re-synchronize itself if the received sequence number, supposedly embedded in the random frame, is outside of a predetermined sequence number window (for example, 255) away from an expected sequence number. Alternatively, the RLP will request a retransmission of all of the data frames between the received and expected sequence numbers. In either case, the RLP error control processes disadvantageously reduce useful data throughput on the channel since most of the available bandwidth is utilized to re-sync the RLP or retransmit numerous data frames.

Therefore, there is a need to more accurately validate data frames at a receiver in a communication system, to thereby reduce the occurrence of such RLP error control processes and correspondingly increase channel bandwidth efficiency over conventional techniques.

## SUMMARY OF THE INVENTION

The present invention provides a method and apparatus for maximizing throughput of a data call in a wireless communication system in which data is transmitted from a wireless station, such as a mobile station, on multiple assigned channels in accordance with a known transmission standard, such as IS-95B. In one embodiment, the multiple assigned channels include a fundamental channel and at least one supplemental channel. Data is formatted into variable rate data frames and transmitted on the fundamental and supplemental channels. A wireless receiver, such as a base station, receives the multiple assigned channels. The wireless receiver demodulates and decodes data frames associated with each of the multiple assigned channels. The wireless receiver determines a likely initial data rate for each demodulated and decode data frame. The wireless receiver correlates all of the likely data rates, by comparison to one another and to a relevant transmission protocol standard, to determine a maximum likelihood combination of data rates. The maximum likelihood combination of data rates

includes a maximum likelihood data rate corresponding to each likely data rate. Decoded data frames are invalidated and erased when the likely data frame rates do not match corresponding maximum likelihood data rates.

Features and Advantages

The present invention overcomes the above mentioned problems and represents an improvement over known rate determination and data validation techniques in a wireless data communication receiver.

The present invention accurately determines a variable transmitted data rate for packet data traffic at a wireless receiver without embedding rate information into the transmitted data.

The present invention advantageously reduces a likelihood that invalid data will be declared valid at the wireless receiver during both periods of data transmission and black-outs. More specifically, the present invention enhances the accuracies of rate determination and data validation at the receiver, and results in an increase in a traffic channel bandwidth efficiency over conventional techniques.

In a communication system including fundamental and supplemental channels operating in accordance with IS-95B, the present invention improves the accuracies of rate determination and data validation on the fundamental channel using supplemental channel signal quality measurements.

## BRIEF DESCRIPTION OF THE FIGURES

The foregoing and other features and advantages of the invention will be apparent from the following, more particular description of the exemplary embodiments of the invention, as illustrated in the accompanying drawings.

FIG. 1 is a block diagram of an exemplary digital communications system 100 in which the present invention can be implemented.

FIG. 1A is an illustration of an exemplary transmit timing diagram of an FCCH and an exemplary transmit timing diagram of a concurrently assigned SCCH.

FIG. 2 is a block diagram of an exemplary transmit channel processor and a block diagram of an exemplary receive channel processor from FIG. 1.

FIG. 3 is an illustration of an exemplary high-level method of determining a maximum likelihood combination of rates used for validating decoded frames at a receiver of FIG. 1.

FIG. 4 is an illustration of a method corresponding to an exemplary embodiment of the present invention, wherein a receiver of FIG. 1 receives IS-95B reverse-link traffic channels.

FIG. 5 is an illustration of three exemplary timing diagrams (a), (b), and (c) corresponding respectively to an FCCH and two assigned SCCHs, and used to illustrate the method of FIG. 4.

FIG. 6 is a block diagram of an exemplary computer system on which the present invention can be implemented.

## DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

FIG. 1 is a block diagram of an exemplary digital communications system 100 in which the present invention can be implemented. In an exemplary embodiment, system 100 is a CDMA cellular telephone system. However, it is to be understood that the present invention is applicable to other types of communication systems such as personal communications systems (PCS), wireless local loop, private branch exchange (PBX) or other known systems. The present invention is also applicable to systems using other well known transmission modulation schemes such as TDMA. System 100 includes a wireless transmitter 110 and a wireless receiver 120, each of which can be part of a base station (also known as a cell-site) or a mobile station. Communication from transmitter 110 to receiver 120 when receiver 120 is disposed in a mobile

station is known as the "forward link," and communication from transmitter 110 to receiver 120 when receiver 120 is disposed in a base station is known as the "reverse link." In the exemplary embodiment, transmitter 110 is disposed in a wireless station, such as the mobile station, and receiver 120 is disposed in the base station. Also, transmitter 110 and receiver 120 operate in accordance with IS-95B. The exemplary CDMA system operating in accordance with IS-95B allows for data communications between users over terrestrial links. The exemplary embodiment also applies to a CDMA system operating in accordance with International Telecommunications Union wireless data communication standards for third generation, International Mobile Telecommunications (IMT-2000).

Exemplary transmitter 110 includes a controller 130 for controlling the operation of transmitter 110 and for exchanging communication signaling information with receiver 120 to assign and de-assign communication channels during call-setup and tear-down, for example. Transmitter 110 includes a transmit channel processor 132 for performing transmit channel processing for one or more communication channels assigned to transmitter 110.

A data source 134 provides data 136 at variable data rates to transmitter 110. Data 136 can be synchronous or asynchronous packet data, as is known in the art. In turn, transmitter 110 formats data 136 into consecutive, variable rate data frames, each having an exemplary duration of 20 milliseconds. In the exemplary embodiment, an RLP processing component (not shown) at transmitter 110 and operating in accordance with TIA/EIA/IS-707 (referred to as "IS-707"), embeds consecutive frame sequence numbers in consecutive data frames for purposes of error correction and control. Then, transmit channel processor 132 further processes the data frames to prepare the data frames for wireless transmission to receiver 120, as will be further described below.

Transmitter 110 transmits the data frames to receiver 120 on a traffic channel 140 assigned to transmitter 110. In the exemplary embodiment, traffic channel 140 is a reverse link IS-95B traffic channel operating in a accordance

with the HSPD Feature of IS-95B. The IS-95B reverse link traffic channel 140 includes a fundamental code channel (FCCH) F, and can include up to seven additional supplemental code channels (SCCHs) $S_0$, $S_1$, $S_2$, $S_3$, $S_4$, $S_5$, $S_6$. The FCCH is a variable rate channel capable of operating at data frame rates (also referred to herein as "rates") including an FCCH full rate, a half rate, a quarter rate, and an eighth rate. The FCCH can carry data 136 from data source 134 and signaling information. Each of the assigned SCCHs $S_0$-$S_6$ can operate at only an SCCH full rate when data is to be transmitted and at a zero rate during DTX periods when no data is available to be transmitted. Under IS-95B, SCCHs $S_0$-$S_7$ can only transmit (at the SCCH full rate) when the FCCH is concurrently transmitting at the FCCH full rate. The present invention takes advantage of this IS-95B traffic channel restriction to improve the accuracies of determining FCCH frame rates and validating received data frames, as will be further described below.

In accordance with IS-95B, the above mentioned rates fall into two categories, namely, a first rate set RS1, and a second rate set RS2. RS1 includes the following rates:

1) FCCH rates of 9600 bps (the RS1 FCCH full rate), 4800 bps, 2400 bps, or 1200 bps; and

2) SCCH rates of 9600 bps (the RS1 SCCH full rate) or zero bps.

On the other hand, RS2 includes the following rates:

1) FCCH rates of 14,000 bps, 7200 bps, 3600 bps, and 1800 bps; and

2) SCCH rates of 14,000 bps or zero bps. It is to be understood that the present invention is applicable to communication systems having a greater or lesser number of data frame rates.

Still with reference to FIG. 1, receiver 120 includes a controller 150 for controlling the receiver and for exchanging signaling information with transmitter 110 to assign and de-assign traffic channels. Receiver 120 also includes a receive channel processor 152 for receiving traffic channel 140 and for processing received data frames so as to recover packet data 154, corresponding to packet

data 136, at transmitter 110. Receiver 120 delivers packet data 154 to a data sink 160. In the exemplary embodiment, receiver 120 and transmitter 110 both implement complementary RLP layers in accordance with IS-707. Controller 150 can include one or more controllers, and can encompass one or more processing functions of receive channel processor 152.

The above mentioned channel transmission requirements of the IS-95B HSPD feature are illustratively depicted in FIG. 1A. An exemplary transmit timing diagram (a) of the FCCH F and an exemplary transmit timing diagram (b) of a concurrently assigned SCCH $S_i$, are depicted in FIG. 1A. Timing diagram (a) is a plot of the FCCH transmitted rate (Rate) versus time, and timing diagram (b) is a plot of the SCCH $S_i$ transmitted rate (Rate) versus time.

Referring to timing diagram (a), transmitter 110 transmits on the FCCH at the full, quarter, half, eighth, and full rates during consecutive portions 172, 174, 176, 178 and 180 of the timing diagram. A time interval 182 represents the duration of a single transmitted data frame, such as 20ms.

Referring to timing diagram (b), transmitter 110 concurrently transmits on SCCH $S_i$ at the SCCH full rate during portions 190 and 192 respectively coinciding with portions 172 and 180 of timing diagram (a), in accordance with IS-95B. Conversely, transmitter 110 transmits on SCCH $S_i$ at a zero rate (that is, transmitter 110 does not transmit) during portion 194 of timing diagram (b) coinciding with portions 174-178 of timing diagram (a). Portion 194 of timing diagram (b) corresponds to a black-out or DTX period on SCCH $S_i$. Also, it is to be understood the FCCH frames can be transmitted at the FCCH full rate while the SCCH $S_i$ is at the zero rate.

FIG. 2 is a block diagram of exemplary transmit channel processor 132 of transmitter 110 and a block diagram of exemplary receive channel processor 152 of receiver 120. In transmit channel processor 132, a variable rate data framer 206 receives variable rate data 136, frames the variable rate data into variable data rate frames (also referred to herein as "frames"), and provides the frames to a cyclic redundancy code and tail bit generator 208, as applicable under IS-95B (for

example, only 9600 and 4800 bps FCCH frames and 9600 SCCH frames receive CRCs under IS-95B RS1). CRC generator 208 generates a set of CRC bits, such as 12 CRC bits, to provide for error detection at receiver 120. In addition, generator 208 appends a sequence of tail bits to each frame. In the exemplary embodiment, generator 208 generates the CRC and tail bits in accordance with IS-95B. Generator 208 provides a data frame to an encoder 210 for encoding the data as symbols for error correction and detection at receiver 120. In the exemplary embodiment, encoder 210 is a convolutional encoder. Encoder 210 provides encoded symbols to an interleaver 212. Interleaver 212 reorders the encoded symbols in accordance with a predetermined interleaving format. In the exemplary embodiment, interleaver 212 is a block interleaver, which is known in the art.

Interleaver 212 provides a reordered data frame to a modulator 214 for modulating the data frame for transmission. In the exemplary embodiment, modulator 214 is a CDMA modulator. Modulator 214 provides a modulated data frame to a transmitter module 216. Transmitter module 216 up-converts and amplifies the up-converted signal for transmission via an antenna 218. Transmitter module 216 transmits data frames to receiver 120 on traffic channel 140.

Receiver 120 receives traffic channel 140 via an antenna 220. Antenna 220 provides the received traffic channel to a plurality of parallel receive channel processors $152_1$-$152_n$. Each of receive channel processors $152_1$-$152_n$ is assigned by receiver controller 150 to perform receive channel processing on a corresponding one of the received traffic channels F and $S_0$-$S_n$ (also referred to herein as "F-$S_n$"). For example, receive channel processor $152_1$ can be assigned to the FCCH, while the next receive channel processor $152_2$ can be assigned to SCCH $S_0$, and so on. In this manner, receive processing for any one of the received channels F-$S_n$ can be performed independently of the receive processing for any of the other received traffic channels.

Receive channel processor $152_1$ performs receive channel processing as is now described. Receive antenna 220 provides received traffic channel 140 to a receiver module 222. Receiver module 222 down converts and amplifies the received traffic channel and provides a down converted and amplified received traffic channel to a demodulator 224, which demodulates the received channel. In the exemplary embodiment, demodulator 224 is a CDMA demodulator. In another embodiment, each of receive channel processors $152_1$-$152_n$ can share a single demodulator. Demodulator 224 provides a demodulated signal, namely, demodulated data frames, to de-interleaver 228. De-interleaver 228 re-orders demodulated data frame symbols in accordance with a predetermined format, as is known in the art.

De-interleaver 228 provides a re-ordered data frame to a decoder 230 for decoding the data frame. In the case where receive channel processor $152_1$ is assigned to the FCCH, decoder 230 is preferably a multi-rate Viterbi decoder capable of decoding FCCH full rate, half rate, quarter rate and eighth rate received data frames associated with the FCCH, as is known in the art. In the case where receive channel processor $152_1$ is assigned to an SCCH, $S_i$, decoder 230 need only decode full rate data frames since SCCH $S_i$ can operate at only the SCCH full rate or the zero rate. As mentioned above, although the transmitted date frame rate can change on a frame by frame basis, rate information is typically not included in each transmitted data frame. Therefore, receiver 120 determines the transmitted rate for each received data frame to accurately decode and validate the data frame.

The decoding and CRC checking processes for a received FCCH frame are now described. In the exemplary embodiment, decoder 230 decodes symbols in the received FCCH frame for each of the four possible transmitted rates (that is, the FCCH full, half, quarter, and eighth rates) so as to provide four separately decoded frames, each of which is provided to a CRC check detector 232. Using conventional techniques, CRC check detector 232 determines whether the CRC bits for each of the four decoded frames are correct. CRC check detector 232

performs a CRC check for the CRC bits in each of the four decoded frames to determine at which of the full, half, quarter, or eighth rates the currently received frame was transmitted. As a result, in one embodiment, CRC check detector 232 provides four check bits, $C_1$, $C_2$, $C_4$, $C_8$, where the subscripts "1", "2", "4", and "8" respectively corresponding to the full rate, half rate, quarter rate, and eighth rate, and where a binary value of "1" for a given CRC check bit can indicate that the CRC check bits passed, while a binary value of "0" can indicate that the CRC bits failed.

In addition, decoder 230 provides decoded frame data to a Symbol Error Rate (SER) detector 234. Specifically, SER detector 234 receives decoded frame bits and an estimate of the received symbol data from decoder 230. As is known, SER detector 234 re-encodes and re-decodes the decoded bits, and compares them to the estimate of the received symbol data from decoder 230. The SER is a count of the number of discrepancies between the re-encoded symbol data and the received symbol data. Therefore, SER detector 234 generates four SER values: $SER_1$, $SER_2$, $SER_3$, and $SER_4$.

Furthermore, decoder 230 provides information to a Yamamoto check detector 236 for providing a confidence metric based on the difference between the selected path through a trellis and the next closest path through the trellis. The Yamamoto quality metric is well known in the art, and is further described, for example, in U.S. Pat. Nos. 5,710,784 and 5,872,775. While the CRC check is dependent on the bits in each of the four decoded frames, the Yamamoto check is dependent on the decoding process of receiver 120. Yamamoto detector 136, similar to detectors 232 and 234, provides four Yamamoto values for each of the four possible rates: $Y_1$, $Y_2$, $Y_4$, and $Y_8$. Although detectors 232, 234, 236 are shown as separate elements, the detectors can be incorporated within the hardware and/or software processes of decoder 230.

Receive channel processor $152_1$ collectively provides the CRC check bits, SER values, and Yamamoto values from respective detectors 232, 234 and 236 to controller or control processor 150 as a data frame quality metric signal $240_1$.

Data frame quality metric signal $240_1$ is indicative of the quality (and thus validity) of decoded data corresponding to the data frame. Using data frame quality metric signal $240_1$, control processor 150 determines at which of the four rates the currently received FCCH data frame was transmitted. For example, in the exemplary embodiment, the control processor selects a rate corresponding to a passed CRC and a favorable SER value.

Receive channel processor $152_1$ also provides a decoded frame signal $242_1$ to the control processor. Decoded frame signal $242_1$ includes each of the separately decoded frames corresponding to the four different frame rates. Decoded frame signal $242_1$ can be provided to a decoded data memory buffer so as to be accessible to the control processor.

The decoding and CRC checking processes performed on a received SCCH frame are similar to those processes described above for a received FCCH frame, as is now described. In the case where a receive channel processor (such as receive channel processor $152_2$) is assigned to SCCH $S_i$, the associated decoder 230 decodes each received data frame at only the SCCH full rate. In this case, the assigned receive channel processor provides a single decoded SCCH data frame to control processor 150. Also, the receive channel processor provides the associated data frame quality metrics (for instance, the CRC, SER and Yamamoto values) associated with the decoded SCCH frame to control processor 150. Thus, in the case where multiple receive channel processors $152_1$-$152_n$ respectively process multiple receive channels F, $S_0$-$S_n$, the receive channel processors respectively provide data frame quality metrics signals $240_1$-$240_n$ and decoded frame signals $242_1$-$242_n$ to the control processor.

High-Level Method

Receiver controller 150 uses the above described signal quality metrics signals $240_1$-$240_n$ to initially determine current FCCH and SCCH frame rates and to initially validate the associated, decoded FCCH and SCCH frames. The

present invention then refines and thus improves the accuracy of such initial determinations, as is further described below.

FIG. 3 is an illustration of an exemplary high-level method 300 of determining a maximum likelihood combination of rates used for validating decoded frames at receiver 120, according to the present invention. Method 300 advantageously improves the likelihood of providing only valid received frames to subsequent processing stages such as the RLP processing layer and/or data sync 160. In doing so, method 300 reduces RLP error processing and correspondingly increases useful traffic channel bandwidth over other known methods, such as, for example, methods using only the above mentioned initial determinations.

Method 300 begins at a step 305 when transmitter 110 transmits data frames on multiple assigned traffic channels F-$S_n$. At a next step 310, receiver 120 receives traffic channels F-$S_n$. At a next step 315, receiver 120 demodulates, de-interleaves and decodes each of the received channels F-$S_n$ as described in connection with FIG. 2.

At a next step 320, a rate for each of the received channels F-$S_n$ is initially determined independent of the other received channels. Each determined or detected rate can be considered a "likely" rate because it may be incorrect if, for example, errors have corrupted the corresponding transmitted frame. In the exemplary embodiment, the likely rate for each SCCH is determined to be the SCCH full rate when the CRC check bits pass and the SER values are favorable for the decoded SCCH frame. When the likely rate is equal to the SCCH full rate, the associated SCCH decoded frame is assumed valid. On the other hand, when the likely rate is determined to be the zero rate, the associated SCCH data frame is assumed invalid.

In the exemplary embodiment, the likely rate for the FCCH is determined based on CRC check bits $C_1$, $C_2$, $C_4$, $C_8$, and SER values $SER_1$, $SER_2$, $SER_3$, and $SER_4$, provided that CRC check bits are available. Specifically, the likely FCCH rate is determined to be the one of the four possible rates corresponding to the one of the four decoded frames having a passing CRC and a favorable SER value.

The decoded frame associated with the selected likely rate is initially assumed valid.

At a next step 325, all of the likely rates determined at step 320 are correlated to produce a Maximum Likelihood (ML) combination of rates for the received traffic channels. The ML combination of rates includes an ML rate corresponding to each likely rate. Each such ML rate can be a probabilistically more accurate estimate of the transmitted rate than is the corresponding likely rate. This is because each likely rate is determined independent of the other traffic channels, whereas the ML rate is determined by correlating all of the independent likely rates. Correlating the independent likely rates adds relevant cross-channel rate information, such as traffic channel interdependencies, to each of the ML rate determinations, to thereby produce a probabilistically better rate estimate.

The correlation includes a comparison of each likely rate to each of the other likely rates. In addition, the correlation can include a comparison of the likely rates to a relevant set of rules, such as the traffic channel transmission requirements for the particular standard (for example, IS-95B) under which the traffic channels were transmitted. Such a comparison adds further relevant information to the process of generating the ML rates. A correlation in accordance with the exemplary embodiment is further described below in connection with FIG. 4.

At a next step 330, one or more of the likely rates determined at step 320 are compared or matched against corresponding ML rates in the ML combination of rates to determine whether to invalidate any of the decoded frames (such as the decoded FCCH frame) initially assumed valid in previous step 320.

Then, all of the decoded frames confirmed as valid in step 330 are provided to the next level of processing, such as the RLP processing layer and/or data sync 160. On the other hand, data frames invalidated at step 325 (and previous step 320) are "erased," that is, such invalidated frames are not provided

to the next level of processing. For example, the FCCH frame and one or more SCCH frames may be invalidated at step 330, based on the results from step 325.

Exemplary Method Embodiment

FIG. 4 is an illustration of a method 400 corresponding to the exemplary embodiment of the present invention, wherein receiver 120 receives reverse-link traffic channels operating in accordance with IS-95B. The principles embodied in exemplary method 400 also apply to any wireless data communication system operating in accordance with IMT-2000. The method steps of FIG. 4 are first described below, and then, a rationale supporting the method steps is provided. Steps 305, 310, 315, and 320 described above in connection with FIG. 3 are collectively represented in a single initial step 405 of method 400.

Next, at a decision step 410, it is determined whether the likely FCCH rate is at the FCCH full rate. If the likely FCCH rate is at the FCCH full rate, the decoded FCCH frame is assumed valid for use at the next processing stage, and flow proceeds to a step 415.

At step 415, decoded frames associated with received SCCHs are validated based on the respective likely rates of the decoded frames, as follows. First, the likely rates for the SCCHs (that is, the likely rate of each SCCH frame transmitted concurrently with the FCCH frame) are determined as described above. Then, decoded SCCH frames associated with full rates and zero rates are respectively assumed valid and invalid. Invalid SCCH frames are erased.

On the other hand, if at step 410 it is determined that the FCCH rate is at other than full rate, then flow proceeds to a next decision step 420. At decision step 420 it is determined whether at least two of the SCCHs have likely rates equal to the SCCH full rate. If at least two of the SCCHs have likely rates equal to the SCCH full rate, then flow proceeds to a step 425 where the decoded FCCH data frame (determined to be at other than the FCCH full rate at step 410) is

assumed invalid and erased. The SCCHs are validated in accordance with their respective likely rates as described in connection with step 415.

On the other hand, if at step 420 it is determined that less than two of the SCCHs are at the SCCH full rate, then flow proceeds to a step 430 where all of the concurrently received SCCH decoded frames are erased.

Decisional Analysis

The decisional logic embodied in method 400 is supported by a combination of the IS-95B requirements described above and by a probability analysis now described. The probability analysis considers two relevant probabilities. A first relevant probability $P_e$ arises when an FCCH frame is transmitted at the FCCH full rate. In this case there is a finite probability, $P_e$, that the likely FCCH rate initially determined at step 410 will be erroneous, that is, the likely rate may be determined to be a rate other than the full rate (such as a half, quarter or eighth rate). This finite probability $P_e$ is referred to as the "rate determination error for a full rate frame". The probability of detecting a full rate frame as other than a full rate frame, that is, the "probability of a rate determination error for a full rate frame" can be determined from Table 1 below. Table 1 is an excerpt from the TIA/EIA-IS-98B "Recommended Minimum Performance Standards for Dual Mode Wideband Spread Spectrum Cellular Mobile Stations" (referred to herein as "IS-98B"). Table 1 tabulates for the FCCH the minimum probabilities of rate determination error for IS-95B Rate Set 1 (RS1) and Rate Set 2 (RS2) full rate frames. RS1 full rate frames are assumed for the present discussion.

| FCCH Rate | Min. Probability of Rate Detection Error at 1% FER (obtained from IS-98B) | |
|---|---|---|
| | *RS1 Full* | *RS2 Full* |
| *Half* | $1.67 \times 10^{-5}$ | $1.67 \times 10^{-5}$ |
| *Quarter* | $1.41 \times 10^{-4}$ | $2.38 \times 10^{-4}$ |
| *Eighth* | $1.73 \times 10^{-4}$ | $2.73 \times 10^{-4}$ |

Table 1: Minimum Probabilities of Rate Determination Errors

Table 1 includes a first column listing FCCH rates, a second column listing error probabilities for RS1 FCCH full rate frames, and a third column listing error probabilities for RS2 FCCH full rate frames. Table 1 includes three rows respectively corresponding to half, quarter and eighth rates. The first row indicates the probability of erroneously detecting a full rate frame as a half rate frame. Similarly, the second row indicates the probability of erroneously detecting a full rate frame as a quarter rate frame, and so on.

The total minimum probability of a rate determination error ($P_e$) for detecting an RS1 FCCH full rate frame as other than a full rate frame is the addition of the error probabilities, from Table 1, of detecting the frame rate as one of the other three frame rates. In other words, the probability of erroneously detecting an FCCH frame transmitted at the full rate as other than the full rate is given by:

$$P_e = 1.67 \times 10^{-5} + 1.41 \times 10^{-4} + 1.73 \times 10^{-4} = 3.31 \times 10^{-4}$$

A second probability, $P_c$, relates to erroneously detecting an invalid received SCCH frame as a valid frame, for example, during a DTX period. As mentioned above, a transmitted SCCH data frame includes a 12 bit CRC. When the CRC passes at receiver 120, the corresponding SCCH frame is assumed valid.

It is to be understood that SER can also used for supplemental rate decisions, but that it is ignored here to simplify this probability analysis. Invalid frames can be received, for example, during a DTX period, when transmitted frames are corrupted with noise, or when transmitted frames are substantially attenuated during transmission. In such circumstances, there is the finite probability $P_c$ of detecting a valid CRC at receiver 120 even though invalid data is being received and demodulated. The random probability $P_c$ of a 12 bit CRC matching any random bit sequence at receiver 120 is $2.4x10^{-4}$. Further, assuming SCCH channels are statistically independent from each other for the purpose of calculating such a random probability, then the random probability $P_{cc}$ of two SCCHs both passing CRCs is given by:

$$P_{cc} = P_c \text{ x } Pc, \text{ where } P_c = 2.4x10^{-4}$$
$$\text{therefore } P_{cc} = 2.4x10^{-4} \text{ x } 2.4x10^{-4} = 5.96x10^{-8}$$

A comparison between $P_{cc}$ and $P_e$ reveals that $P_{cc} << P_e$, by several orders of magnitude. Since SCCH data frames can only be transmitted (at the SCCH full rate) when FCCH data frames are transmitted at the FCCH full rate under IS-95B, the probabilistic comparison $P_{cc}$ vs. $P_e$ definitively suggests the following conclusion: when a FCCH frame is detected at a rate other than the full rate (for example, at the half, quarter or eighth rate) and at the same time or concurrently (that is, for the same 20 ms frame interval) at least two SCCH data frames associated with two SCCHs are detected at the full rate, it is much more likely than not that the FCCH non full rate determination is erroneous and that the FCCH data frame was actually transmitted at the full rate. In other words, the initial FCCH non full-rate determination is most likely wrong, and therefore should be overruled.

Under such circumstances, it is likely the FCCH data frame is corrupted (or a DTX period is in progress) and probabilities dictate that it is safer to invalidate and erase the FCCH frame than it is to provide such a corrupted frame

to the RLP. Method 400 thus improves FCCH rate detection during HSPD calls by filtering-out invalid FCCH data frames in accordance with the result of the above described correlation between all of the received traffic channel rates, and the further comparison of the rates against the IS-95B transmission requirements.

The above described probabilistic comparison $P_{cc}$ vs. $P_e$ definitively suggests the FCCH non full-rate determination should be overruled when at least two SCCHs are at the full-rate. On the other hand, when only one SCCH is determined to be at the full rate, a relevant probabilistic comparison $P_c$ ($2.4 \times 10^{-4}$) vs. $P_c$ ($3.31 \times 10^{-4}$) is much less definitive since $P_c$ and $P_e$ are substantially the same, that is, within an order of magnitude of one another. Relative to the earlier probability comparison, this comparison suggests it is just as likely the FCCH data frame was transmitted at the FCCH full rate as it was not transmitted at the FCCH full rate when only one SCCH channel is detected at the SCCH full rate. Under such conditions, probability does not justify overruling a determination that the FCCH is not full-rate based on a single SCCH channel being full rate.

Therefore, in the exemplary embodiment, when the FCCH rate is not full rate and only one SCCH is full rate, the SCCH frame is invalidated/erased while the FCCH data frame is assumed valid and provided to the next processing stage. This approach is taken because experience has shown erasure of a valid SCCH data frame is less harmful than providing an invalid SCCH frame to the RLP.

Method 400 is now illustrated with reference to FIG. 5. FIG. 5 is an illustration of exemplary timing diagrams (a), (b) and (c) corresponding respectively to the FCCH and two assigned SCCHs. In diagrams (a), (b) and (c), the timing waveforms in solid line represent transmitted frame rates. At receiver 120, the detected rates (that is, the determined likely rates) are in accordance with the transmitted rates, except during a first frame interval 505 and a second frame interval 510 (depicted in timing diagram (a)), where respective erroneous likely rates 505' (timing diagram (c)) and 510' (timing diagram (a)) are depicted in dotted line.

During interval 505, while the FCCH rate is at the half rate, $SCCH_2$ is erroneously determined to be at the SCCH full rate (that is, the SCCH likely rate is equal to the SCCH full rate). Such a condition is not allowed under IS-95B. In this situation, method 400 invalidates and erases a decoded $SCCH_2$ frame associated with interval 505 in favor of the FCCH half rate detected during the same time interval.

During interval 510, while the FCCH is erroneously determined to be at the FCCH half rate, at least two concurrent SCCH full rate frames are detected, namely, full rate frames for $SCCH_1$ and $SCCH_2$. Such a condition is not allowed under IS-95B. In this situation, method 400 invalidates and erases the decoded FCCH frame in favor of the two SCCH full rate frames.

Table 2 below provides an exemplary illustration of the operation of method 400. Table 2 tabulates SCCH and FCCH frame erasure decisions in accordance with method 400 when up to four SCCH are assigned and received at receiver 120. The legend or key for interpreting Table 2 is as follows:

F = Full Rate; and

!F = not full rate (that is, Quarter, Half or Eighth Rate);

| Fund | S1 | S2 | S3 | S4 | S5 | S6 | S7 | Action |
|---|---|---|---|---|---|---|---|---|
| !F | F | | | | | | | Erase S1 |
| !F | F | F | | | | | | Erase F |
| !F | F | !F | | | | | | Erase S1 |
| !F | !F | F | | | | | | Erase S2 |
| !F | !F | !F | | | | | | Erase S1,S2 |
| !F | F | F | F | | | | | Erase F |
| !F | F | F | !F | | | | | Erase F, S3 |
| !F | F | !F | F | | | | | Erase F,S2 |
| !F | F | !F | !F | | | | | Erase S1 |
| !F | !F | F | F | | | | | Erase F,S1 |
| !F | !F | F | !F | | | | | Erase S2 |
| !F | !F | !F | F | | | | | Erase S3 |
| !F | !F | !F | !F | | | | | Erase S1,S2,S3 |
| !F | F | F | F | F | | | | Erase F |
| !F | F | F | F | !F | | | | Erase F, S4 |
| !F | F | F | !F | F | | | | Erase F, S3 |
| !F | F | F | !F | !F | | | | Erase F,S3,S4 |
| !F | F | !F | F | F | | | | Erase F,S2 |
| !F | F | !F | F | !F | | | | Erase F,S2, S4 |
| !F | F | !F | !F | F | | | | Erase F,S2,S3 |
| !F | F | !F | !F | !F | | | | Erase S1 |
| !F | !F | F | F | F | | | | Erase F,S1 |
| !F | !F | F | F | !F | | | | Erase F,S1,S4 |
| !F | !F | F | !F | F | | | | Erase F,S1,S3 |
| !F | !F | F | !F | !F | | | | Erase S2 |
| !F | !F | !F | F | F | | | | Erase F,S1,S2 |
| !F | !F | !F | F | !F | | | | Erase S3 |
| !F | !F | !F | !F | F | | | | Erase S4 |
| !F | !F | !F | !F | !F | | | | Erase S1-S4 |

Table 2 : Example of new algorithm

Receiver 120 can perform specific features of the present invention using receiver controllers, which in effect comprise a computer system. Although communication-specific hardware can be used to implement the present invention, the following description of a general purpose computer system is provided for completeness. The present invention is preferably implemented in software. Alternatively, the invention may be implemented using hardware or a combination of hardware and software. Consequently, the invention may be implemented in a computer system or other processing system.

An example of such a computer system 600 is shown in FIG. 6. In the present invention, for example, the above described methods or processes execute on computer system 600. The computer system 600 includes one or more processors, such as processor 604. The processor 604 is connected to a communication infrastructure 606 (for example, a bus or network). Various software implementations are described in terms of this exemplary computer system. After reading this description, it will become apparent to a person skilled in the relevant art how to implement the invention using other computer systems and/or computer architectures.

Computer system 600 also includes a main memory 608, preferably random access memory (RAM), and may also include a secondary memory 610. The secondary memory 610 may include, for example, a hard disk drive 612 and/or a removable storage drive 614, representing a floppy disk drive, a magnetic tape drive, an optical disk drive, etc. The removable storage drive 614 reads from and/or writes to a removable storage unit 618 in a well known manner. Removable storage unit 618, represents a floppy disk, magnetic tape, optical disk, etc. which is read by and written to by removable storage drive 614. As will be appreciated, the removable storage unit 618 includes a computer usable storage medium having stored therein computer software and/or data.

In alternative implementations, secondary memory 610 may include other similar means for allowing computer programs or other instructions to be loaded into computer system 600. Such means may include, for example, a removable

storage unit 622 and an interface 620. Examples of such means may include a program cartridge and cartridge interface (such as that found in video game devices), a removable memory chip (such as an EPROM, or PROM) and associated socket, and other removable storage units 622 and interfaces 620 which allow software and data to be transferred from the removable storage unit 622 to computer system 600.

Computer system 600 may also include a communications interface 624. Communications interface 624 allows software and data to be transferred between computer system 600 and external devices. Examples of communications interface 624 may include a modem, a network interface (such as an Ethernet card), a communications port, a PCMCIA slot and card, etc. Software and data transferred via communications interface 624 are in the form of signals 628 which may be electronic, electromagnetic, optical or other signals capable of being received by communications interface 624. These signals 628 are provided to communications interface 624 via a communications path 626. Communications path 626 carries signals 628 and may be implemented using wire or cable, fiber optics, a phone line, a cellular phone link, an RF link and other communications channels.

In this document, the terms "computer program medium" and "computer usable medium" are used to generally refer to media such as removable storage drive 614, a hard disk installed in hard disk drive 612, and signals 628. These computer program products are means for providing software to computer system 600.

Computer programs (also called computer control logic) are stored in main memory 608 and/or secondary memory 610. Computer programs may also be received via communications interface 624. Such computer programs, when executed, enable the computer system 600 to implement the present invention as discussed herein. In particular, the computer programs, when executed, enable the processor 604 to implement the process of the present invention. Accordingly, such computer programs represent controllers of the computer

system 600. By way of example, in a preferred embodiment of the invention, the processes performed by receiver controller 150 can be performed by computer control logic. Where the invention is implemented using software, the software may be stored in a computer program product and loaded into computer system 600 using removable storage drive 614, hard drive 612 or communications interface 624.

In another embodiment, features of the invention are implemented primarily in hardware using, for example, hardware components such as application specific integrated circuits (ASICs). Implementation of the hardware state machine so as to perform the functions described herein will be apparent to persons skilled in the relevant art(s).

While various embodiments of the present invention have been described above, it should be understood that they have been presented by way of example, and not limitation. It will be apparent to persons skilled in the relevant art that various changes in form and detail can be made therein without departing from the spirit and scope of the invention.

The present invention has been described above with the aid of functional building blocks illustrating the performance of specified functions and relationships thereof. The boundaries of these functional building blocks have been arbitrarily defined herein for the convenience of the description. Alternate boundaries can be defined so long as the specified functions and relationships thereof are appropriately performed. Any such alternate boundaries are thus within the scope and spirit of the claimed invention. One skilled in the art will recognize that these functional building blocks can be implemented by discrete components, application specific integrated circuits, processors executing appropriate software and the like or any combination thereof. Thus, the breadth and scope of the present invention should not be limited by any of the above-described exemplary embodiments, but should be defined only in accordance with the following claims and their equivalents.

What is Claimed is:

1. A method of maximizing throughput of a data call in a wireless communication system in which data is transmitted from a wireless station on multiple assigned channels, comprising the steps of:

a. receiving the multiple assigned channels;

b. demodulating and decoding each of the multiple assigned channels;

c. determining a likely data rate of each of the multiple assigned channels; and

d. correlating all of the likely data rates to determine one or more Maximum Likelihood (ML) data rates each corresponding to a likely data rate.

2. The method of claim 1, further comprising the step of

e. invalidating data associated with one of the multiple assigned channels when the likely data rate and a corresponding ML data rate of the one of the multiple assigned channels do not match.

3. The method of claim 1, wherein the multiple assigned channels include a fundamental channel and a supplemental channel, and

wherein data can be transmitted at a first data rate on the fundamental channel, and

wherein data can be transmitted at a second data rate on the supplemental channel only when data is being transmitted at the first data rate on the fundamental channel, and

wherein it is more likely than not that data is being transmitted at the first data rate on the fundamental channel when a plurality of supplemental channels have likely data rates equal to the second data rate,

the method further comprising invalidating and erasing demodulated and decoded data associated with the fundamental channel when

a) the fundamental channel does not have a likely data rate equal to the first data rate, and

b) the plurality of supplemental channels have likely data rates equal to the second data rate.

4. The method of claim 3, wherein the multiple assigned channels collectively form an IS-95B reverse-link traffic channel, and

wherein the first data rate corresponds to a fundamental channel full rate and the second rate corresponds to a supplemental channel full rate,

the method further comprising invalidating and erasing demodulated and decoded data associated with each of the plurality of supplemental channels when

a) the fundamental channel does not have a likely data rate equal to the fundamental channel full rate, and

b) only one of the plurality of supplemental channels has a likely data rate equal to the supplemental channel full rate.

5. The method of claim 4, further comprising the step of providing non-invalidated data to a radio link protocol processing layer.

6. The method of claim 1, wherein the multiple assigned channels include a fundamental channel and a supplemental channel, and

wherein data can be transmitted at a first non-zero data rate on the fundamental channel, and

wherein data can be transmitted at a second non-zero data rate on the supplemental channel only when data is being transmitted at the first data rate on the fundamental channel, and

wherein it is approximately equally likely that data is being transmitted and that data is not being transmitted at the first data rate on the fundamental channel when only one of a plurality of supplemental channels has a likely data rate equal to the second data rate,

the method further comprising invalidating and erasing demodulated and decoded data associated with each of the plurality of supplemental channels when

a) the fundamental channel does not have a likely data rate equal to the first data rate, and

b) only one of the plurality of supplemental channels has a likely data rate equal to the second data rate.

7. The method of claim 1, wherein the data transmitted on the multiple assigned channels is formatted into data frames, and wherein step (b) comprises the steps of:

demodulating the data frames to produce demodulated data frames; and

de-interleaving the demodulated data frames to produce de-interleaved data frames;

8. The method of claim 7, further comprising the steps of:

decoding the de-interleaved data frames to produce decoded data frames; and

generating a signal quality signal indicative of a signal quality for each of the decoded data frames.

9. The method of claim 8, likely data rate of each of the decoded data frames based on a corresponding signal quality metric signal.

10. The method of claim 8, wherein each of the data frames includes a Cyclic Redundancy Code (CRC), and wherein the generating step comprises at least one of:

generating a CRC for each of the decoded data frames; and

generating a Symbol Error Rate (SER) for each of the decoded data frames.

11. The method of claim 10, wherein step (c) comprises determining a likely data rate of each of the data frames on each of the multiple assigned channels based on at least one of a CRC and an SER for each of the decoded data frames.

12. Apparatus for maximizing throughput of a data call in a wireless communication system in which data is transmitted by a wireless station to a receiver on multiple assigned channels, comprising:

receiving means for receiving the multiple assigned channels;

demodulating means and decoding means for respectively demodulating and decoding each of the multiple assigned channels;

determining means for determining a likely data rate of each of the multiple assigned channels; and

correlating means for correlating all of the likely data rates to determine a maximum likelihood combination of data rates.

13. The apparatus of claim 12, wherein the maximum likelihood combination of data rates includes a maximum likelihood data rate corresponding to each said likely data rate, the apparatus further comprising

invalidating means for invalidating data associated with one of the multiple assigned channels when the likely data rate of the one multiple assigned channel as determined by the determining means fails to match a corresponding maximum likelihood data rate determined by the correlating means.

14. The apparatus of claim 12, wherein the multiple assigned channels include a fundamental channel and a supplemental channel, and

wherein data can be transmitted at a first data rate on the fundamental channel, and

wherein data can be transmitted at a second data rate on the supplemental channel only when data is being transmitted at the first data rate on the fundamental channel, and

wherein it is more likely than not that data is being transmitted at the first data rate on the fundamental channel when a plurality of supplemental channels have likely data rates equal to the second data rate,

the apparatus further comprising means for invalidating and erasing demodulated and decoded data associated with the fundamental channel when

a) the fundamental channel does not have a likely data rate equal to the first data rate, and at the same time,

b) the plurality of supplemental channels have likely data rates equal to the second data rate.

15. The apparatus of claim 14, wherein the multiple assigned channels collectively form an IS-95B reverse-link traffic channel, and

wherein the first data rate corresponds to a fundamental channel full rate and the second rate corresponds to a supplemental channel full rate,

the apparatus further comprising means for invalidating and erasing demodulated and decoded data associated with the plurality of supplemental channels when

a) the fundamental channel does not have a likely data rate equal to the fundamental channel full rate, and

b) only one of the plurality of supplemental channels has a likely data rate equal to the supplemental channel full rate.

16. The apparatus of claim 15, further comprising a radio link protocol processing layer and means for providing non-invalidated data to the radio link protocol processing layer.

17. The apparatus of claim 13, wherein the multiple assigned channels include a fundamental channel and a supplemental channel, and

wherein data can be transmitted at a first non-zero data rate on the fundamental channel, and

wherein data can be transmitted at a second non-zero data rate on the supplemental channel only when data is being transmitted at the first data rate on the fundamental channel, and

wherein it is approximately equally likely that data is being transmitted and that data is not being transmitted at the first data rate on the fundamental channel when only one of a plurality of supplemental channels has a likely data rate equal to the second data rate,

the apparatus further comprising means for invalidating and erasing demodulated and decoded data associated with the plurality of supplemental channels when

a) the fundamental channel does not have a likely data rate equal to the first data rate, and

b) only one of the plurality of supplemental channels has a likely data rate equal to the second data rate.

18. The apparatus of claim 13, wherein the data transmitted on the multiple assigned channels is formatted into data frames, and wherein:

the demodulating means includes means for demodulating the data frames to produce demodulated data frames; and

the de-interleaving means includes means for de-interleaving the demodulated data frames to produce de-interleaved data frames.

19. The apparatus of claim 18, wherein

the decoding means include means for decoding the de-interleaved data frames to produce decoded data frames; and

generating means for generating a signal quality signal indicative of a signal quality for each of the decoded data frames.

20. The apparatus of claim 19, wherein the determining means includes means for determining a likely data rate of each of the decoded data frames based on a corresponding signal quality metric signal;

21. The apparatus of claim 19, wherein each of the data frames includes a Cyclic Redundancy Code (CRC), and wherein the generating means comprises at least one of:

means for generating a CRC for each of the decoded data frames; and

means for generating a Symbol Error Rate (SER) for each of the decoded data frames.

22. The apparatus of claim 21, wherein the determining means determines a likely data rate of each of the data frames on each of

the multiple assigned channels based on at least one of a CRC and an SER for each of the decoded data frames.

23. A computer program product comprising computer usable media having computer readable program code means embodied in the media for causing application programs to execute on a computer processor in a wireless communication device to maximize throughput of a data call in a wireless communication system in which data is transmitted by a wireless station to the wireless communication device on multiple assigned channels, the wireless communication device including receiving means for receiving the multiple assigned channels, and demodulating and decoding means for demodulating and decoding each of the multiple assigned channels, the computer readable program code means comprising:

a first computer readable program code means for causing the processor to determine a likely data rate of each of the multiple assigned channels; and

a second computer readable program code means for causing the processor to correlate all of the likely data rates to determine a maximum likelihood combination of data rates.

24. The computer program product of claim 23, further comprising a third computer readable program code means for causing the processor to invalidate data associated with one of the multiple assigned channels when the likely data rate of the one multiple assigned channel fails to match a corresponding maximum likelihood data rate.

25. The computer program product of claim 23, wherein the multiple assigned channels include a fundamental channel and a supplemental channel, and

wherein data can be transmitted at a first data rate on the fundamental channel, and

wherein data can be transmitted at a second data rate on the supplemental channel only when data is being transmitted at the first data rate on the fundamental channel, and

wherein it is more likely than not that data is being transmitted at the first data rate on the fundamental channel when a plurality of supplemental channels have likely data rates equal to the second data rate,

the computer program product further comprising a third computer readable program code means for causing the processor to invalidate and erase demodulated and decoded data associated with the fundamental channel when

a) the fundamental channel does not have a likely data rate equal to the first data rate, and

b) the plurality of supplemental channels have likely data rates equal to the second data rate.

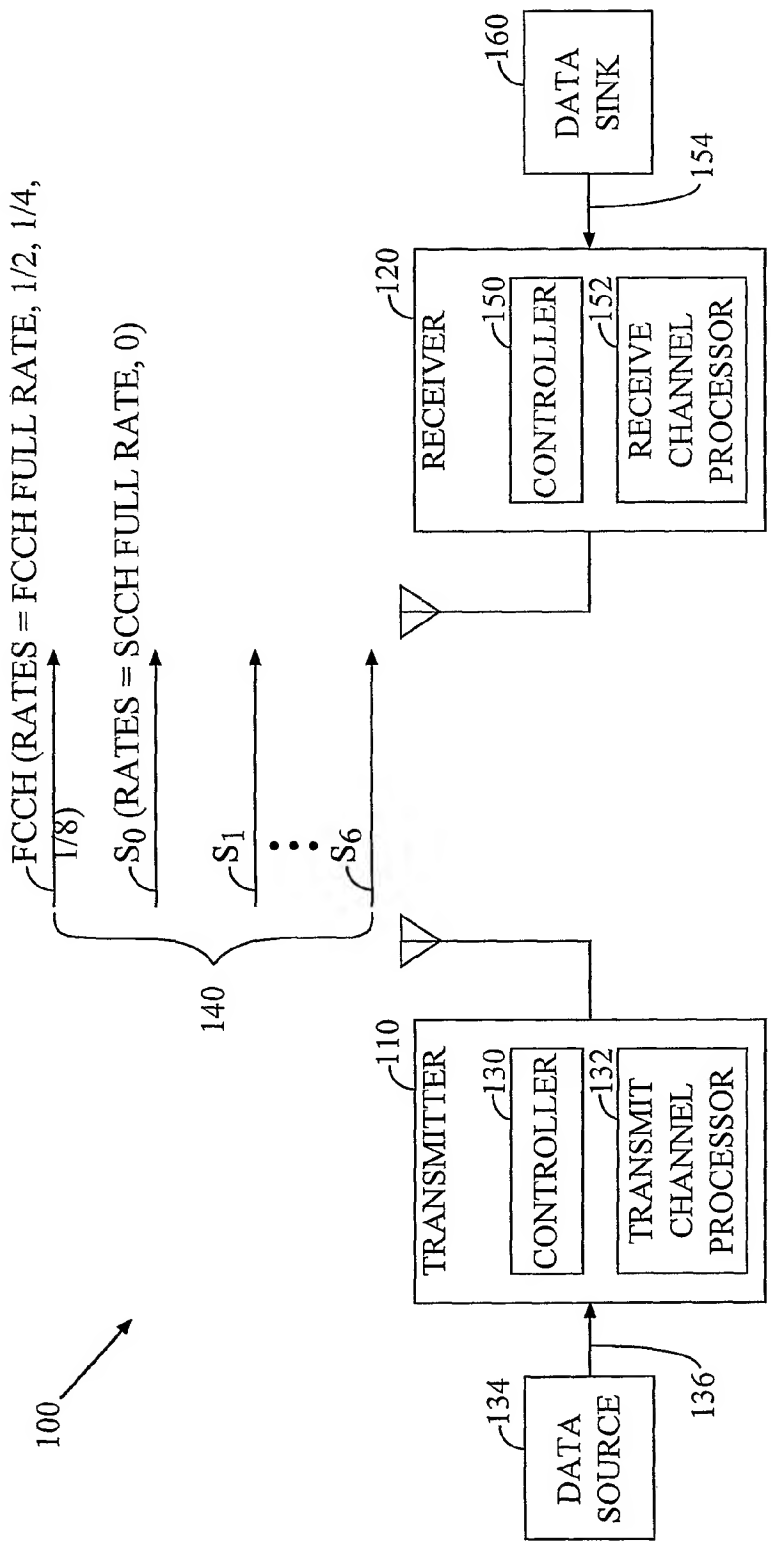
26. The computer program product of claim 25, wherein the multiple assigned channels collectively form an IS-95B reverse-link traffic channel, and

wherein the first data rate corresponds to a fundamental channel full rate and the second rate corresponds to a supplemental channel full rate,

the computer program product further comprising a fourth computer readable program code means for causing the processor to invalidate and erase demodulated and decoded data associated with each of the plurality of supplemental channels when

a) the fundamental channel does not have a likely data rate equal to the fundamental channel full rate, and

b) only one of the plurality of supplemental channels has a likely data rate equal to the supplemental channel full rate.



100
DATA SOURCE
134
136
TRANSMITTER
110
CONTROLLER
130
TRANSMIT CHANNEL PROCESSOR
132
140
FCCH (RATES = FCCH FULL RATE, 1/2, 1/4, 1/8)
$S_0$ (RATES = SCCH FULL RATE, 0)
$S_1$
$S_6$
RECEIVER
120
CONTROLLER
150
RECEIVE CHANNEL PROCESSOR
152
154
DATA SINK
160


FIG. 1

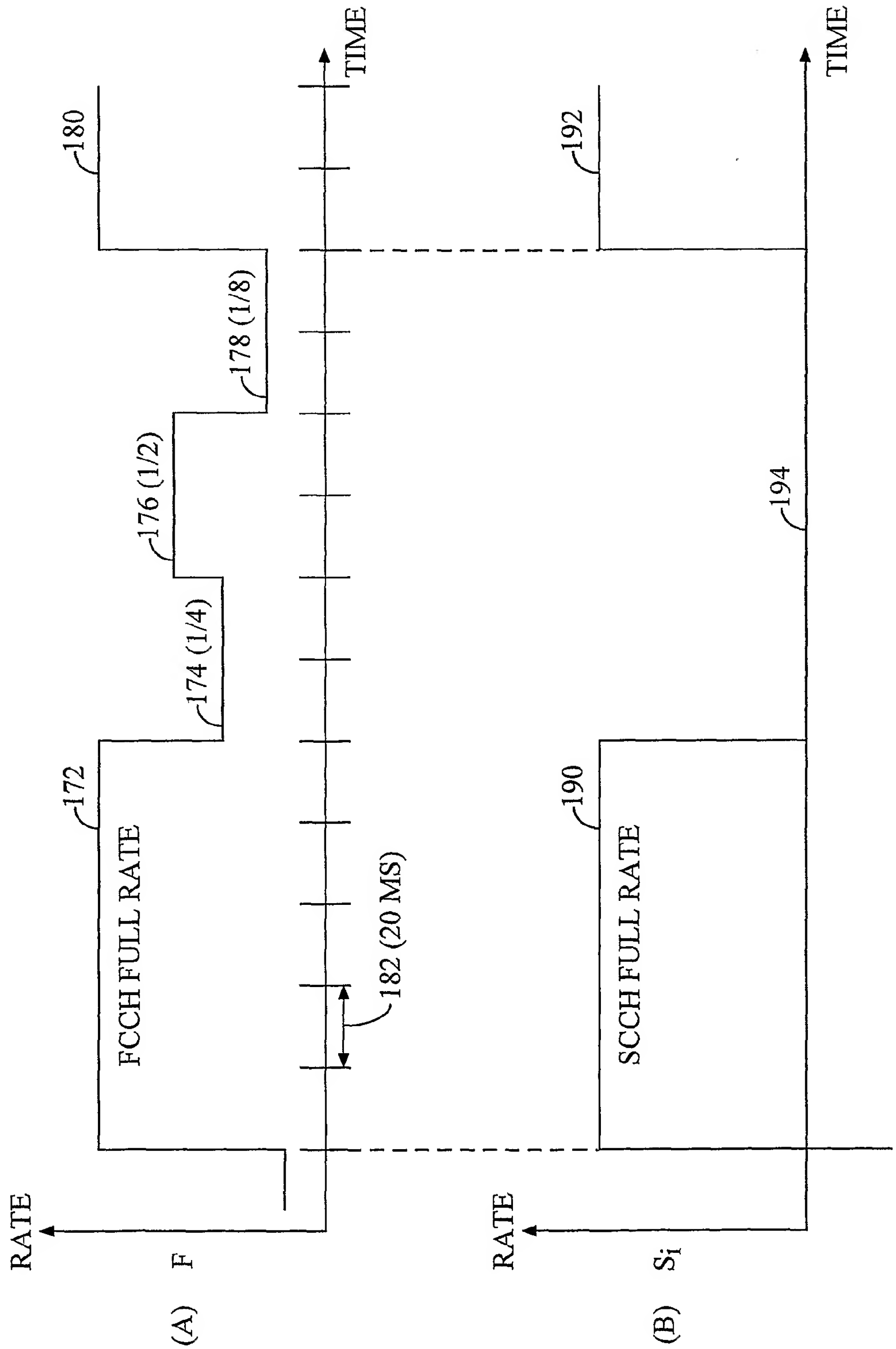

(A)
RATE
F
FCCH FULL RATE
172
174 (1/4)
176 (1/2)
178 (1/8)
180
182 (20 MS)
TIME
(B)
RATE
$S_i$
SCCH FULL RATE
190
192
194
TIME

FIG. 1A

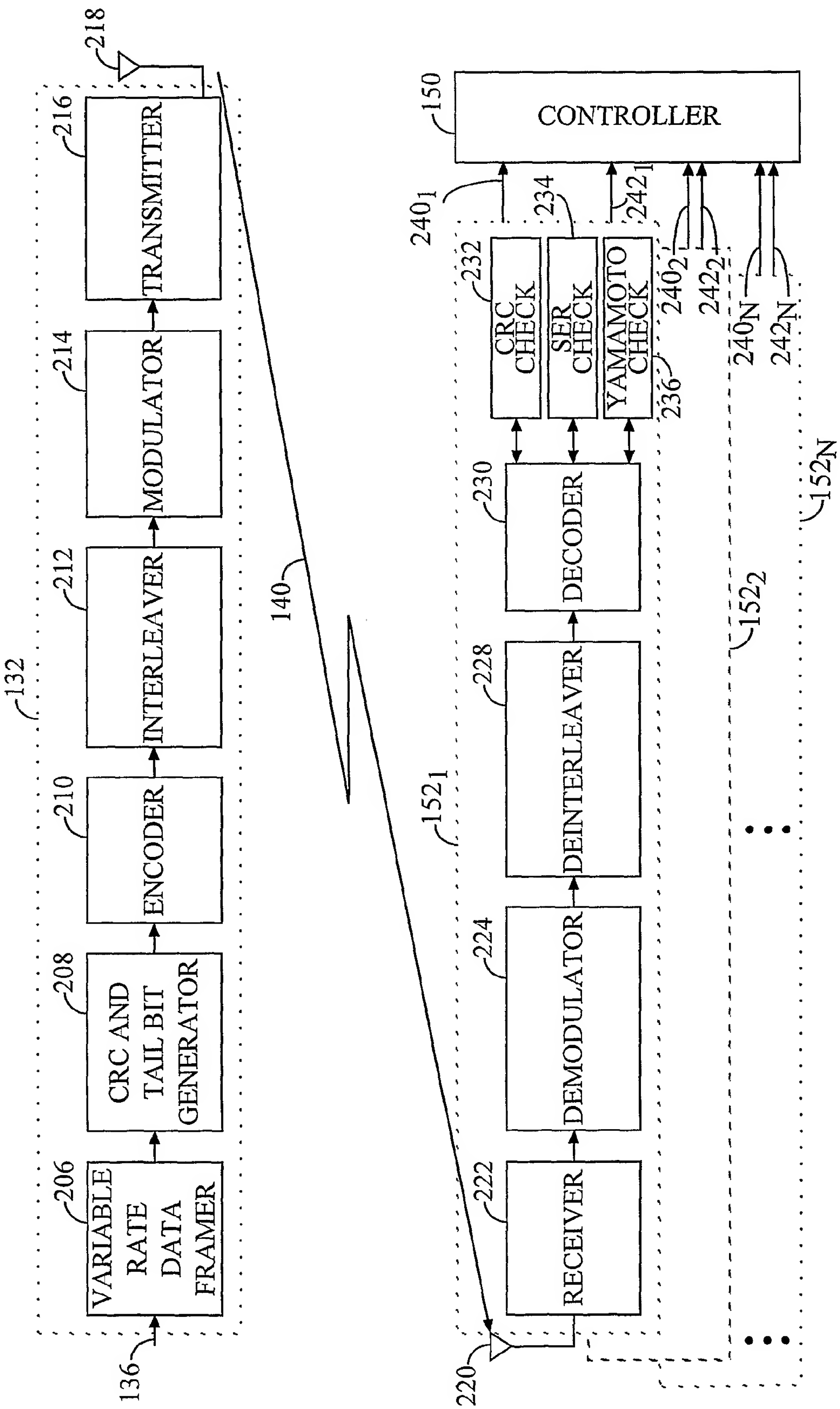

136
206
VARIABLE RATE DATA FRAMER
208
CRC AND TAIL BIT GENERATOR
210
ENCODER
212
INTERLEAVER
214
MODULATOR
216
TRANSMITTER
218
132
140
150
CONTROLLER
220
222
RECEIVER
224
DEMODULATOR
228
DEINTERLEAVER
230
DECODER
232
CRC CHECK
234
SER CHECK
236
YAMAMOTO CHECK
$240_1$
$242_1$
$240_2$
$242_2$
$240_N$
$242_N$
$152_1$
$152_2$
$152_N$

FIG. 2

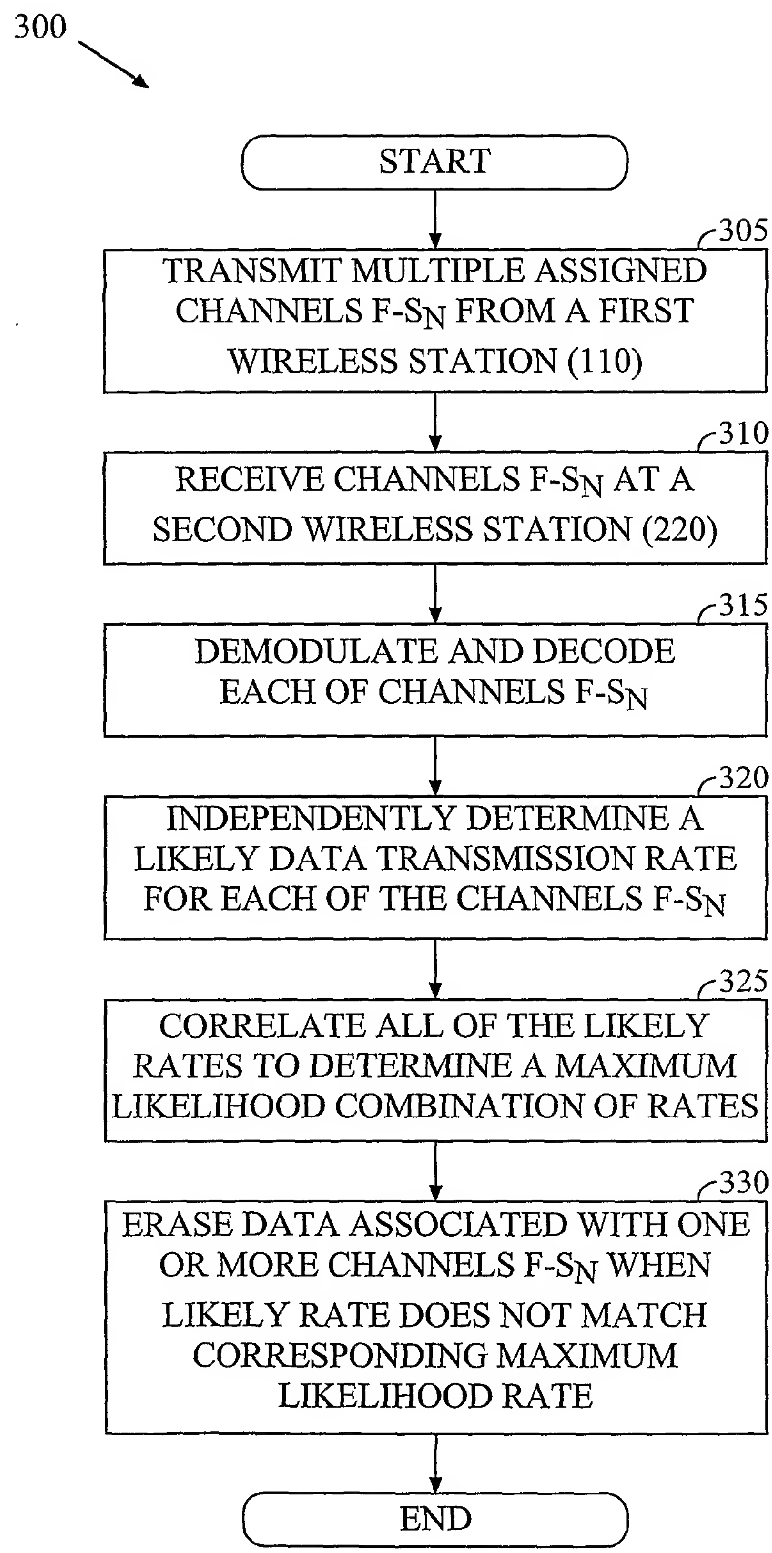

300
START
305
TRANSMIT MULTIPLE ASSIGNED CHANNELS F-$S_N$ FROM A FIRST WIRELESS STATION (110)
310
RECEIVE CHANNELS F-$S_N$ AT A SECOND WIRELESS STATION (220)
315
DEMODULATE AND DECODE EACH OF CHANNELS F-$S_N$
320
INDEPENDENTLY DETERMINE A LIKELY DATA TRANSMISSION RATE FOR EACH OF THE CHANNELS F-$S_N$
325
CORRELATE ALL OF THE LIKELY RATES TO DETERMINE A MAXIMUM LIKELIHOOD COMBINATION OF RATES
330
ERASE DATA ASSOCIATED WITH ONE OR MORE CHANNELS F-$S_N$ WHEN LIKELY RATE DOES NOT MATCH CORRESPONDING MAXIMUM LIKELIHOOD RATE
END

FIG. 3

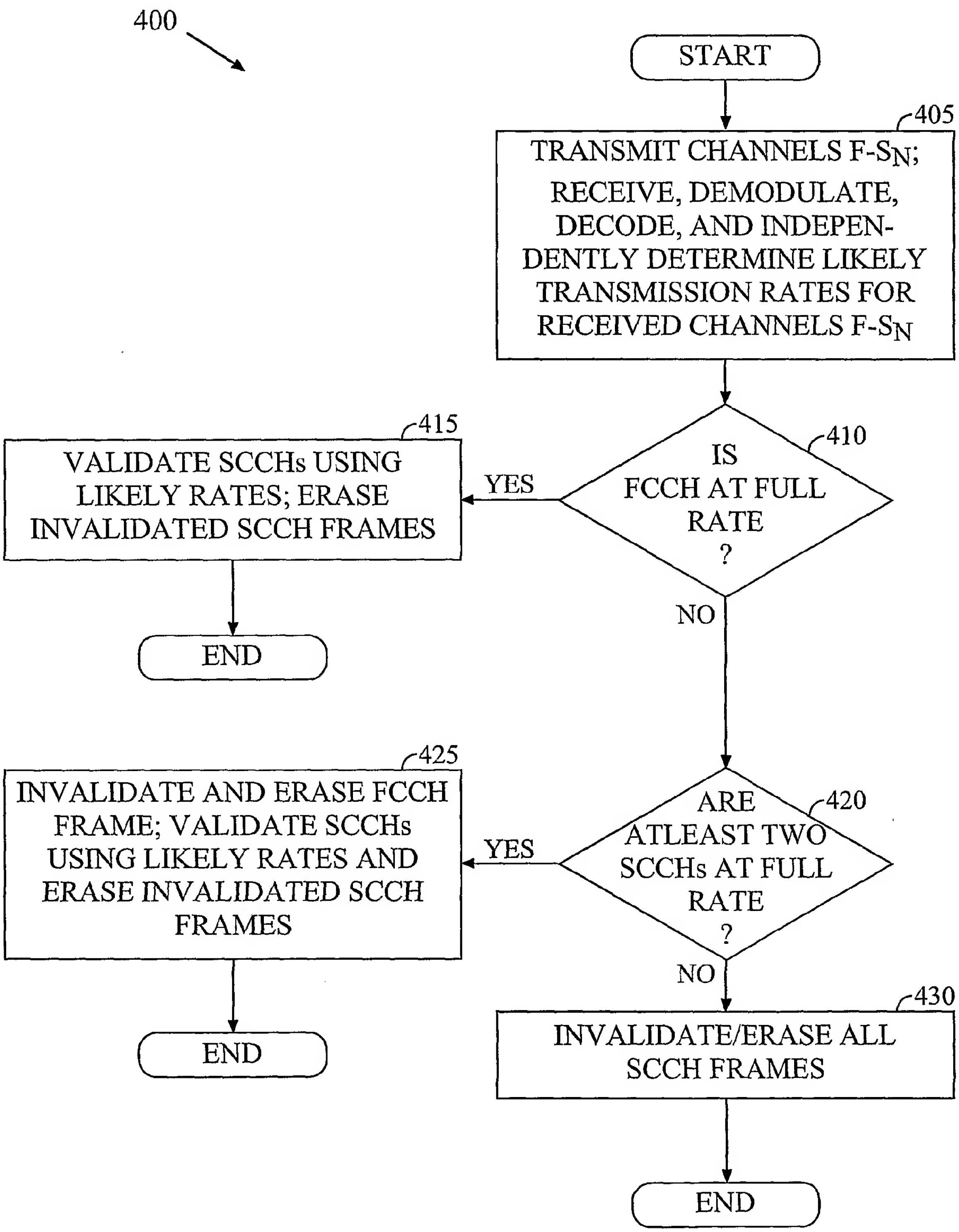

400
START
405
TRANSMIT CHANNELS F-$S_N$; RECEIVE, DEMODULATE, DECODE, AND INDEPENDENTLY DETERMINE LIKELY TRANSMISSION RATES FOR RECEIVED CHANNELS F-$S_N$
410
IS FCCH AT FULL RATE ?
YES
415
VALIDATE SCCHs USING LIKELY RATES; ERASE INVALIDATED SCCH FRAMES
END
NO
420
ARE ATLEAST TWO SCCHs AT FULL RATE ?
YES
425
INVALIDATE AND ERASE FCCH FRAME; VALIDATE SCCHs USING LIKELY RATES AND ERASE INVALIDATED SCCH FRAMES
END
NO
430
INVALIDATE/ERASE ALL SCCH FRAMES
END

FIG. 4

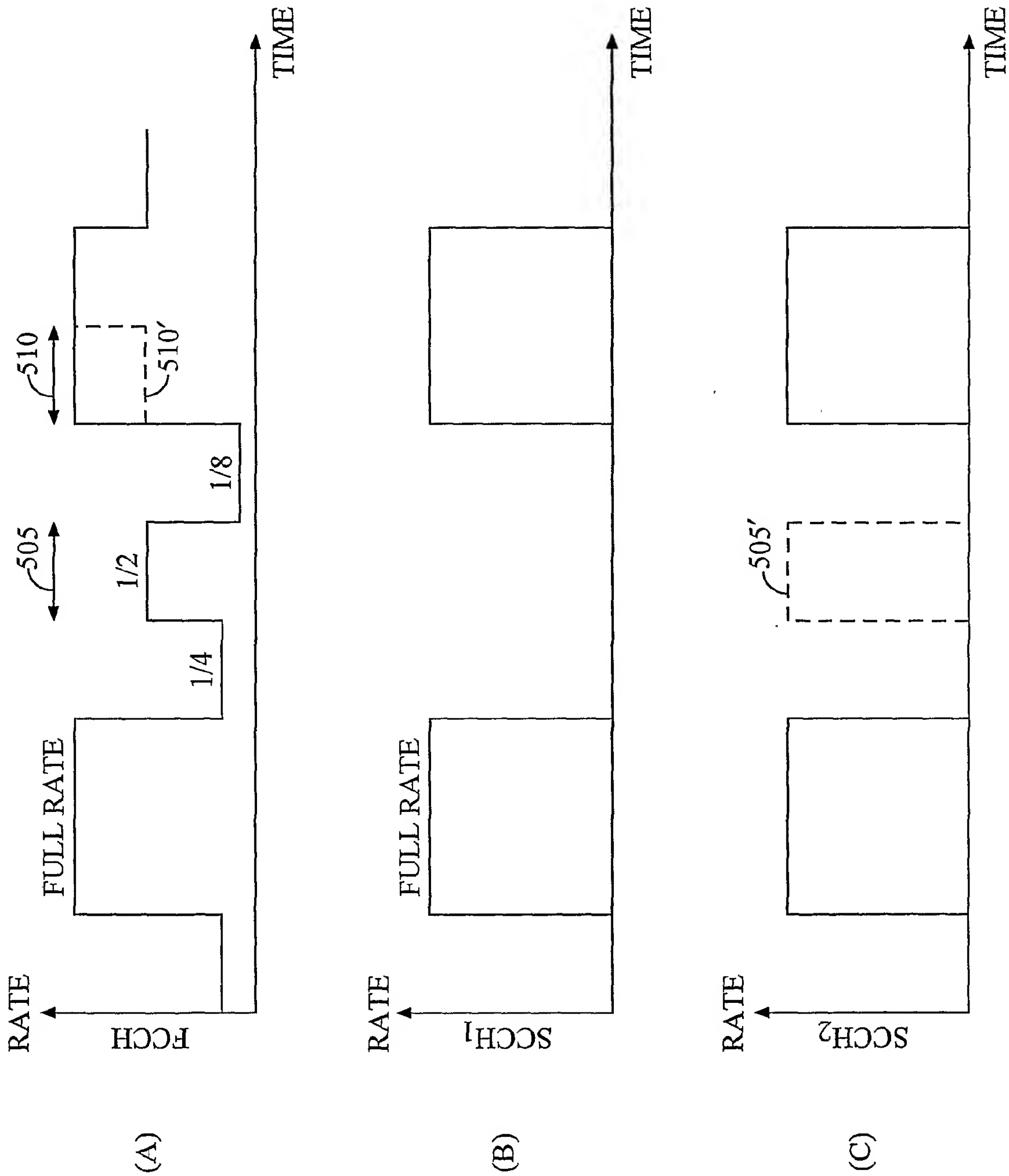

RATE
FCCH
FULL RATE
1/4
505
1/2
1/8
510
510′
TIME
(A)
RATE
SCCH1
FULL RATE
TIME
(B)
RATE
SCCH2
505′
TIME
(C)

FIG. 5

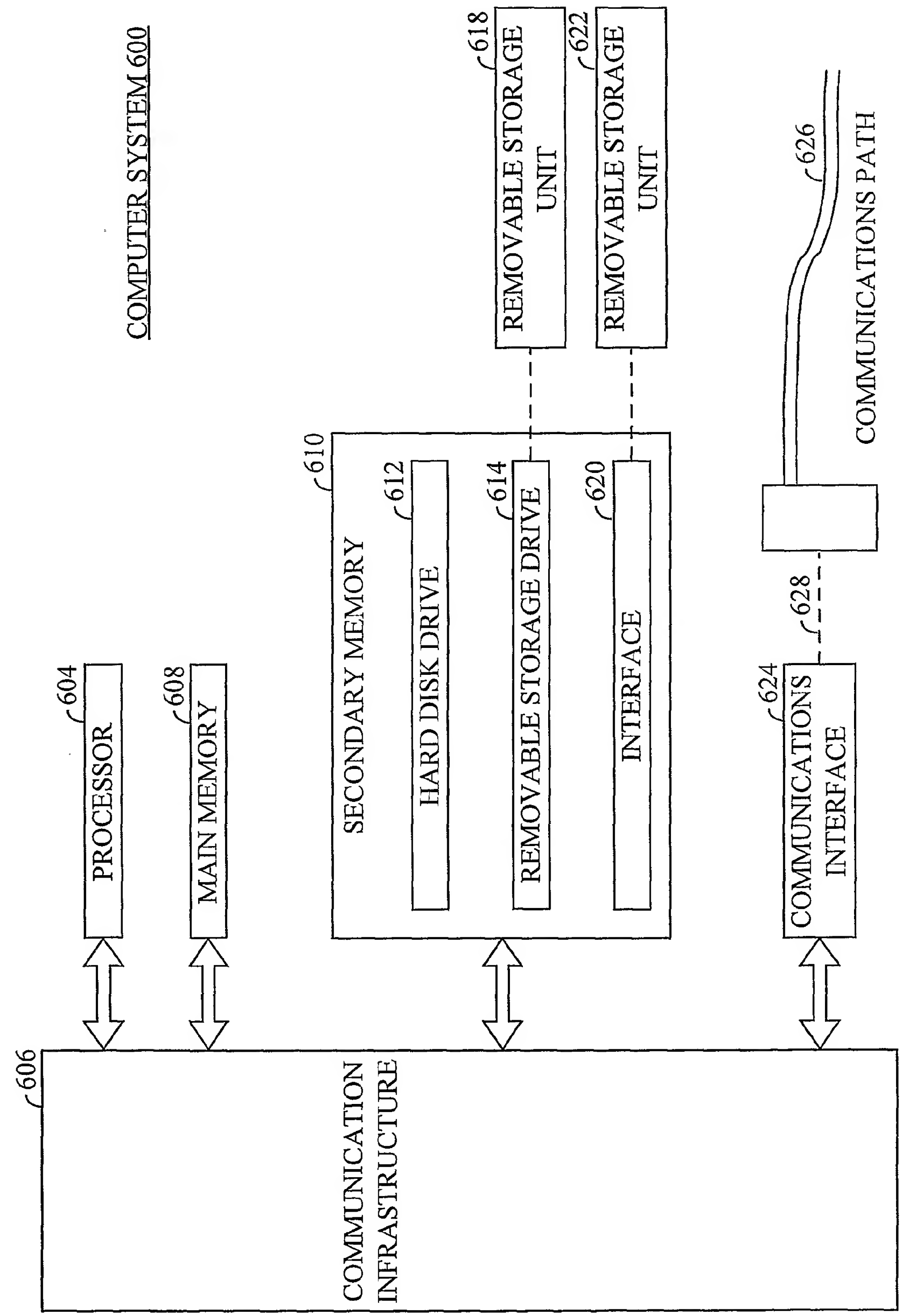

COMPUTER SYSTEM 600
604
PROCESSOR
608
MAIN MEMORY
610
SECONDARY MEMORY
612
HARD DISK DRIVE
614
REMOVABLE STORAGE DRIVE
620
INTERFACE
618
REMOVABLE STORAGE UNIT
622
REMOVABLE STORAGE UNIT
624
COMMUNICATIONS INTERFACE
628
626
COMMUNICATIONS PATH
606
COMMUNICATION INFRASTRUCTURE

FIG. 6

# INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 02/05417

**A. CLASSIFICATION OF SUBJECT MATTER**
IPC 7 H04L25/02

According to International Patent Classification (IPC) or to both national classification and IPC

**B. FIELDS SEARCHED**

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal

**C. DOCUMENTS CONSIDERED TO BE RELEVANT**

| Category ° | Citation of document, with indication, where appropriate, of the relevant passages | Relevant to claim No. |
|---|---|---|
| Y | US 6 108 372 A (LIN YU-CHUAN ET AL) 22 August 2000 (2000-08-22) figures 2,5 | 1,7-12, 23 |
| Y | ANSI/TIA/EIA: "95-B Mobile Station - Base Station Compatibility Standard for Wideband Spread Spectrum Cellular Systems" 1 March 1999 (1999-03-01), TELECOMMUNICATIONS INDUSTRY ASSOCIATION, ARLINGTON XP002206171 page 6-46, line 14-27 footnote on page 6-47 page 7-75, line 6-8 | 1,7-12, 23 |

-/--

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

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"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

| Date of the actual completion of the international search | Date of mailing of the international search report |
|---|---|
| 16 July 2002 | 01/08/2002 |

| Name and mailing address of the ISA | Authorized officer |
|---|---|
| European Patent Office, P.B. 5818 Patentlaan 2<br>NL – 2280 HV Rijswijk<br>Tel. (+31-70) 340-2040, Tx. 31 651 epo nl,<br>Fax: (+31-70) 340-3016 | Horbach, C |

Form PCT/ISA/210 (second sheet) (July 1992)

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

| Category ° | Citation of document, with indication,where appropriate, of the relevant passages | Relevant to claim No. |
|---|---|---|
| A | EP 0 920 160 A (MATSUSHITA ELECTRIC IND CO LTD) 2 June 1999 (1999-06-02)<br>abstract<br>column 4, line 50 -column 5, line 42;<br>figures 4,5<br>----- | 1-26 |

# INTERNATIONAL SEARCH REPORT

ırmation on patent family members

Interr ıal Application No
PCT/US 02/05417

| Patent document cited in search report | | Publication date | Patent family member(s) | | Publication date |
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| EP 0920160 | A | 02-06-1999 | JP | 11163961 A | 18-06-1999 |
| | | | EP | 0920160 A2 | 02-06-1999 |
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(81) Designated States *(national)*: AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO, NZ, OM, PH, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VN, YU, ZA, ZM, ZW.

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**Published:**
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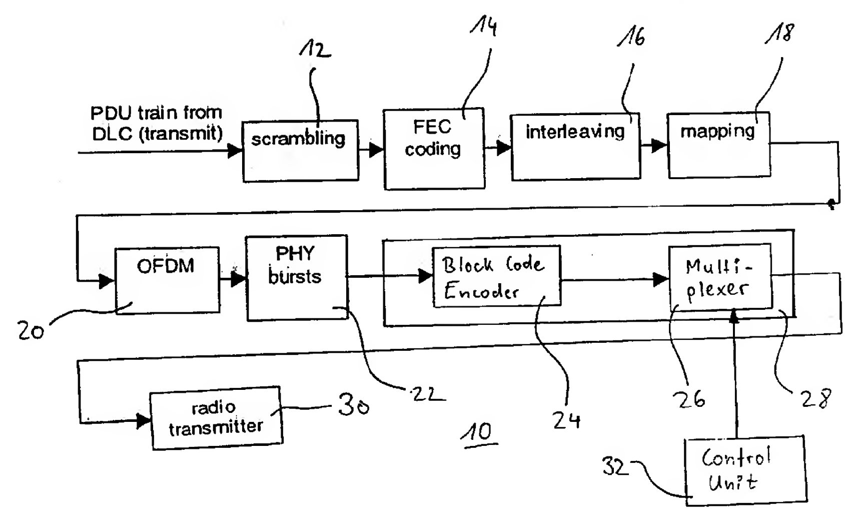
*[Continued on next page]*

(54) **Title:** MULTIPLEXING METHOD IN A MULTI CARRIER TRANSMIT DIVERSITY SYSTEM




(57) **Abstract:** Multiplexing Method in a Multi Carrier Transmit Diversity SystemThe invention relates to a method of multiplexing data words in a multicarrier transmit diversity system. The method comprises the step of generating a plurality of data blocks, each data block comprising data words and each data word containing data symbols derived from a data signal, the step of determining for one or more data blocks in dependence on at least one transmission constraint if the data words of said one or more data blocks are to be multiplexed in the time domain or in the frequency domain and the step of multiplexing the data words of the data blocks in accordance with the determination result.

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— *before the expiration of the time limit for amending the claims and to be republished in the event of receipt of amendments*

*For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.*

# Multiplexing Method in a Multi Carrier Transmit Diversity System

## BACKGROUND OF THE INVENTION

### Technical Field

The present invention relates to the field of transmit antenna diversity and in particular to a method of multiplexing data words in a multi carrier transmit diversity system. The invention also relates to a multiplexer for multiplexing a sequence of data symbols and a demultiplexer for demultiplexing a multiplexed sequence of data symbols.

### Discussion of the Prior Art

Peak transmission rates in wireless communication systems have steadily increased during the last years. However, peak transmission rates are still limited for example due to path loss, limited spectrum availability and fading.

Transmitter diversity is a highly effective technique for combating fading in wireless communications systems. Several different transmit diversity schemes have been proposed. In Li, Y.; Chuang, J.C.; Sollenberger, N.R.: Transmitter diversity for OFDM systems and its impact on high-rate data wireless networks, IEEE Journal on Selec. Areas, Vol. 17, No. 7, July 1999 the transmit diversity schemes of delay, permutation and space-time coding are examplarily described. According to the delay approach, a signal is transmitted from a first transmitter antenna and signals transmitted from further transmitter antennas are delayed versions of the signal from the first transmitter antenna. In the permutation scheme, the modulated signal is transmitted from a first transmitter antenna and permutations of the modulated signal

are transmitted from further transmitter antennas. By means of space-time coding a signal is encoded into several data words and each data word is transmitted from a different transmitter antenna. During transmission the data words are spread (or multiplexed) in the time domain by successively transmitting the data symbols of a data word over a single carrier frequency.

A further transmit diversity scheme for a multicarrier system is space-frequency coding. By means of space-frequency coding a signal is encoded into several data words and each data word is spread (or multiplexed) in the frequency domain by transmitting the data symbols of each data word on orthogonal frequencies, i.e. orthogonal subcarriers. An exemplary scheme for space-frequency coding is described in Mudulodu, S.; Paulraj, A.: A transmit diversity scheme for frequency selective fading channels, Proc. Globecom, San Francisco, pp. 1089-1093, Nov. 2000. According to the multicarrier system described in this paper, the data words relating to an encoded signal are preferably multiplexed in the time domain although orthogonal frequencies are available and the data words could thus also be multiplexed in the frequency domain. This is due to the fact that if multiplexing in the frequency domain is utilized the employed frequencies, i.e. subcarriers, must see the same channel, which may not always be possible in a frequency selective fading channel. However, in case the subcarriers experience the same channel, it is stated that either multiplexing in the time domain or multiplexing in the frequency domain or a combination of the two may be used. By combining multiplexing in the time domain and in the frequency domain the data symbols of a data word are simultaneously multiplexed in the time domain and in the frequency domain. This means that the data word is spread both across time and across frequencies.

Departing from the various transmit diversity schemes hitherto known there is a need for a method of multiplexing data

words in a multicarrier transmit diversity system which can easily be adapted to the specifications of different wireless communications systems. There is also a need for a corresponding multiplexer and a demultiplexer.

BRIEF DESCRIPTION OF THE INVENTION

The existing need is satisfied by a method of multiplexing data words in a multicarrier transmit diversity system which comprises the step of generating a plurality of data blocks, each data block comprising data words and each data word containing data symbols derived from a data signal, the step of determining for one or more of the data blocks in dependence on at least one transmission constraint if the data words of said one of more data blocks are to be multiplexed in a time domain or in a frequency domain and the step of multiplexing the data words of the data blocks in accordance with the result of the determination.

The multiplexing method of the invention is not restricted to a specific transmit diversity scheme as long as the utilized transit diversity scheme enables to generate from a data signal a plurality of data blocks having the above structure. For example, the transmit diversity schemes of block coding and of permutation allow to generate such data blocks. Preferably, the generated data blocks have the structure of a matrix similar to a space-time block code (STBC) matrix. Also, it is not required that the transmit diversity scheme guarantees full transmit diversity. In other words: the invention does not necessitate that each information symbol comprised within the data signal is transmitted from each transmitter antenna. Nonetheless, a preferred embodiment of the invention comprises the feature of full transmit diversity.

Moreover, the invention is not restricted to any number of transmit and receive antennas. Preferably, the number of data words per data block equals the number of transmit antennas

such that each data word of a data block may be transmitted from an individual transmitter antenna. If more than one receive antenna is provided, the receive diversity scheme of maximum-ratio combining can be applied. However, other receive diversity schemes may be used as well.

According to the invention, it is decided on a data block level how the data words are to be multiplexed. The decision on the data block level allows to change the multiplexing domain from one data block to a subsequent data block which is advantageous if one has to cope with specific predefined or varying transmission constraints. Also, the multiplexing method according to the invention can be applied in various wireless communication systems without major changes due to the specific multiplexing flexibility gained by selecting the multiplexing domain on the data block level. The multiplexing domain can be determined for each data block individually or simultaneously for a plurality of data blocks. For example, it can be decided for a sequence of data blocks that all data words comprised within the sequence of data blocks are to be multiplexed in either the time domain or in the frequency domain.

The multiplexing domain is determined by taking into account one or more transmission constraints. For example, the transmission constraints may comprise one or more physical transmission constraints or one or more data-related transmission constraints. It can also comprise both one or more physical transmission constraints and one or more data-related transmission constraints. The physical transmission constraints relate to the physical transmission conditions and can be derived from physical transmission parameters like a channel coherence bandwidth or a coherence time. The data-related transmission constraints relate to system specific constraints regarding for example the employed multiplexing scheme for the data words, the structure of the data signal,

the structure of the data blocks, the structure of the data words or the structure of the data symbols.

The data symbols may be derived from the data signal in various ways dependent on the transmit diversity scheme which is used. If, for example, the transmit diversity scheme of permutation is used, the data symbols contained in the data words are permutations of information symbols comprised within the data signal. As a further example, if the transmit diversity scheme of block coding is used, the data symbols contained in the data words are obtained from the information symbols comprised within the data signal by means of permutation and basic arithmetic operations, such as negation and complex conjugation.

The data signal from which the one or more data blocks are generated can have any format. According to a preferred embodiment, the data signal has the format of a sequence of discrete information symbols. For example, the data signal may have the structure of vectors, each vector comprising a predefined number of information symbols. The nature of the information symbols may depend on the specific wireless communication system in which the multiplexing method according to the invention is used. Many wireless communication systems employ different types of information symbols for different purposes. For example, some wireless communication systems use data signals which comprise a preamble, one or more user data sections or both a preamble and one or more user data sections. Usually, the preamble has a predefined structure and is utilized for purposes like channel estimation, frequency synchronization and timing synchronization.

In the following, several exemplary data-related transmission constraints are described in more detail. According to a first embodiment, the data-related transmission constraint is a predefined number N of data symbols to be comprised within each data word which is to be multiplexed in the time domain.

Usually, the number N of data symbols to be comprised within each data word cannot arbitrarily be chosen because it may depend for example on a code rate, on the condition that the data blocks have to be orthogonal matrices or on the availability of memory resources within the multicarrier transmit diversity system.

When the data words of a specific data block are to be multiplexed in the time domain, the number N of data symbols to be comprised within each data word may represent the number of time slots required for the transmission of a single data word over a single subcarrier. On the other hand, when the data words of a specific data block are to be multiplexed in the frequency domain, the number N of data symbols to be comprised within each data word stands for the number of subcarriers required to transmit a single data word during a single time slot.

Preferably, all data words of an individual data block comprise the same number of data symbols. If the data signal has such a structure that the number of data symbols comprised within each data word of a specific data block equals the predefined number N of data symbols, the data words of this data block may be multiplexed in the time domain. Otherwise, i.e. if the data signal has such a structure that the number of data symbols comprised within each data word of a specific data block does not equal the predefined number N of data symbols, the data words of this data block may be multiplexed in the frequency domain. Such a distinction will become necessary if the data signal or a portion thereof has a predefined length because the predefined length may imply that the total number $N_D$ of data symbols which corresponds to the predefined length of the data signal or a portion thereof is not an integer multiple of the predefined number N of data symbols which should be comprised within a data word to be multiplexed in the time domain. In such a situation integer multiples of the predefined number N of data symbols are ar-

ranged in data blocks of data words which are multiplexed in a time domain and a remainder $N_R = \mathrm{mod}(N_D/N)$ of data symbols is arranged in a data block with data words which are multiplexed in the frequency domain.

Thus, by combining multiplexing in the time domain and in the frequency domain, data symbol fitting problems resulting from the predefined number N of data symbols to be comprised within each data word which is to be multiplexed in the time domain can be solved. Such data symbol fitting problems may for example become relevant when the data signal or a portion of the data signal has a predefined length because the wireless transmission system necessitates that the preamble portion or the user data portion of a data signal comprises a certain number of information symbols. Thus the data words of all data blocks except for the last data block are multiplexed in the time domain and the data words of the last data block are either multiplexed in the time domain or in the frequency domain depending on whether or not the data words of the last data block contain a number of data symbols which equals the predefined number N of data symbols.

So far the data-related transmission constraint of a predefined number N of data symbols to be comprised within each data word has been illustrated. According to a second embodiment, the data signal may comprise one or more periodic structures and the data related transmission constraint may be a preservation of the periodic structures such that the periodic structures are still periodic on a receiver side. The one or more periodic structures may be comprised within a preamble of the data signal, for example in the form of two or more identical preamble information symbols. Periodic structures are advantageous because they allow the use of synchronization algorithms with comparatively low complexity.

In case of multiplexing data symbols relating to periodic structures in the time domain the periodicity of the periodic

structures may get lost. Therefore, at least the data words of data blocks which relate to the periodic structures or parts of periodic structures are multiplexed in the frequency domain. By multiplexing the data words of these data blocks in the frequency domain it can be ensured that the periodicity of the periodic structures is maintained.

When the data words of data blocks generated from periodic structures or portions thereof are multiplexed in the frequency domain, the data words of data blocks generated from the remaining data signal are preferably multiplexed in the time domain. If, for example, the data words of data blocks generated from a preamble comprising periodic structures are multiplexed in the frequency domain, the data words of data blocks generated from a corresponding user data section may be multiplexed at least partly in the time domain.

Instead of data-related transmission constraints or in addition to data-related transmission constraints physical transmission constraints can be taken into account when deciding if the data words of one or more specific data blocks are to be multiplexed in the time domain or in the frequency domain. According to a preferred embodiment, the decision is made based on simultaneously evaluating a combination of one or more data-related transmission constraints and one or more physical transmission constraints.

The physical transmission constraints may be determined based on at least one of a channel coherence bandwidth

$$B_c \approx 1/\tau_{rms} \quad (1)$$

and a coherence time

$$t_c \approx 1/(2 \cdot f_D) \quad (2)$$

wherein $f_D$ is the doppler frequency and $\tau_{rms}$ is the root mean square of the delay spread of the channel impulse response.

Many transmit diversity schemes require constant or at least approximately constant channel parameters during transmission of one data word. If the data words are to be multiplexed in the frequency domain, a comparatively large coherence band width is required. This means that the relation

$$B_C \gg N/T \qquad (3)$$

has to be fulfilled at least approximately, wherein N is the number of data symbols per data word and T is the duration of one of the data symbols, i.e. the duration of one time slot. A comparatively large coherence bandwidth requires that the channel parameters of N adjacent subcarriers have to be almost constant.

On the other hand, if the data words are to be multiplexed in the time domain, a comparatively large coherence time is required. This means that the relation

$$t_C \gg T \cdot N \qquad (4)$$

has to be fulfilled at least approximately. In other words: N subsequent data symbols have to have nearly constant channel parameters, i.e. the channel parameters for a single subcarrier have to remain constant for a period of $N \cdot T$.

The physical transmission constraint may be determined by assessing if one or both of the relations (3) and (4) are fulfilled. Dependent on which of the two relations (3) and (4) is fulfilled best it is decided that the data words of the data blocks are to be multiplexed either in the time domain or in the frequency domain as a general rule. Deviations from this general rule may become necessary due to data-related

transmission constraints. For example, the data symbol fitting problem or the problem encountered with periodic structures may necessitate that although multiplexing in the time domain is generally to be preferred, the data words of at least some data blocks have to be multiplexed in the frequency domain. As a further example, changing transmission conditions may necessitate that the data words of some data blocks have to be multiplexed in the time domain and the data words of other data blocks have to be multiplexed in the frequency domain. As a third example, the data words of data blocks generated from a preamble may be multiplexed in the time domain and the data words of data blocks generated from a user data section may be multiplexed in the frequency domain. Such a combination has the advantage that the above-mentioned data symbol fitting problem, which usually is most relevant for the user data section, can be avoided while the multiplexing in the time domain of the data words of data blocks generated from the preamble allows a good channel estimation.

It was mentioned above that in order to achieve full diversity each information symbol has to be transmitted from each transmitter antenna. A further requirement of full transmit diversity is that the antenna signals are orthogonal to each other. This means that the data symbols have to be modulated onto subcarriers which are orthogonal to each other. However, the invention can also be practiced in case the subcarriers are not orthogonal.

## BRIEF DESCRIPTION OF THE DRAWINGS

Further advantages of the invention will become apparent by reference to the following description of a preferred embodiment of the invention in the light of the accompanying drawings, in which :

Fig. 1 shows a data signal in the form of a physical burst to be processed in accordance with the invention;

Fig. 2 is a block diagram of a transceiver for wireless communication adapted to multiplex data words in accordance with the invention;

Fig. 3 shows several modulation schemes defined in the HIPERLAN/2 standard;

Fig. 4 shows the block code encoder of the transceiver depicted in Fig. 2;

Fig. 5 shows the configuration of a transmit antenna diversity scheme;

Fig. 6 is a schematic diagram of multiplexing data words in the time domain in accordance with the invention; and

Fig. 7 is a schematic diagram of multiplexing data words in the frequency domain in accordance with the invention.

## DESCRIPTION OF PREFERRED EMBODIMENTS

Although the present invention can be used in any multicarrier transmit diversity system which employs a transmit diversity scheme allowing to generate data blocks having a structure as described above, the following description of preferred embodiments is examplarily set forth with respect to a multicarrier system which employs orthogonal frequency division multiplexing (OFDM) and which utilizes block coding for generating data blocks from a data signal.

The exemplary multicarrier system described below is derived from the European wireless local area network (WLAN) standard

high performance radio local area network type 2 (HIPERLAN/2). HIPERLAN/2 systems are intended to be operated in the 5 GHz frequency band. A system overview of HIPERLAN/2 is given in ETSI TR 101 683, Broadband Radio Access Networks (BRAN); HIPERLAN Type 2; System Overview, V1.1.1 (2000-02) and the physical layer of HIPERLAN/2 is described in ETSI TS 101 475; Broadband Radio Access networks (BRAN); HIPERLAN Type 2; Physical (PHY) Layer, V1.1.1 (2000-04). The multicarrier scheme of OFDM, which is specified in the HIPERLAN/2 standard, is very robust in frequency selective environments.

Up to now, the HIPERLAN/2 system and many other wireless communications systems do not support transmit diversity in spite of the fact that transmit diversity would improve the transmission performance and reduce negative effects of fast fading like Rayleigh fading. However, applying standard transmit diversity schemes to multicarrier communications systems may lead to various problems which are hereinafter examplarily described with respect to the HIPERLAN/2 system.

In Fig. 1 a typical physical burst of HIPERLAN/2 is illustrated. The physical burst comprises a preamble consisting of preamble symbols and a user data section consisting of user data symbols. In HIPERLAN/2 five different physical bursts are specified and each kind of physical burst has a unique preamble. However, the last three preamble symbols constitute a periodic structure which is identical for all preamble types. This periodic structure consists of a short OFDM symbol C32 of 32 samples followed by two identical regular OFDM symbols C64 of 64 samples. The short OFDM symbol C32 is a cyclic prefix which is a repetition of the second half of one of the C64 OFDM symbols. The so-called C-preamble depicted in Fig. 1 is used in HIPERLAN/2 for channel estimation, frequency synchronization and timing synchronization. The periodic structure within the C-preamble is necessary in order to allow the use of synchronization algorithms with comparatively low complexity.

The user data section of the physical burst depicted in Fig. 1 comprises a variable number $N_{SYM}$ of OFDM symbols required to transmit a specific protocol data unit (PDU) train. Each OFDM symbol of the user data section consists of a cyclic prefix and a useful data part. The cyclic prefix consists of a cyclic continuation of the useful data part and is inserted before it. Thus, the cyclic prefix is a copy of the last samples of the useful data part. The length of the useful data part is equal to 64 samples and has a duration of 3,2 $\mu$s. The cyclic prefix has a length of either 16 (mandatory) or 8 (optional) samples and a duration of 0,8 $\mu$s or 0,4 $\mu$s, respectively. Altogether, a OFDM symbols thus has a length of either 80 or 72 samples corresponding to a symbol duration of 4,0 $\mu$s or 3,6 $\mu$s, respectively. An OFDM symbol therefore has an extension in the time domain. A OFDM symbol further has an extension in the frequency domain. According to HIPERLAN/2, a OFDM symbol extends over 52 subcarriers. 48 subcarriers are reserved for complex valued subcarrier modulation symbols and 4 subcarriers are reserved for pilots.

From the above it becomes clear that the HIPERLAN/2 physical burst depicted in Fig. 1 has a predefined length both in a time direction and in a frequency direction. Moreover, the physical burst of Fig. 1 comprises a periodic structure. It are among others these features of the physical burst of Fig. 1 which may lead to problems when the HIPERLAN/2 system or a similar wireless communication system has to be adapted to transmit diversity.

For typical HIPERLAN/2 scenarios the above relation (4) is usually fulfilled because the doppler frequency $f_D$ is comparatively low. However, especially in outdoor environments, relatively large delay spreads can occur. Consequently, relation (3) cannot always be fulfilled. Therefore, a transmit diversity scheme like STBC multiplexing in the time domain should generally be a preferred transmit diversity scheme for

a HIPERLAN/2 scenario from the point of view that the channel over one space-time data word should be as constant as possible. However, severe problems arise when STBC is applied to physical bursts having the structure depicted in Fig. 1 or a similar structure.

Both the physical burst and the OFDM symbols comprised therein have predefined dimensions in the time domain and in the frequency domain. Concurrently, STBC requires that each STBC data word has a predetermined length N. Thus, data unit fitting problems arise if the dimension of e.g. an OFDM symbol of the preamble or of the user data section cannot be mapped on an integer multiple of the length of one STBC data word. Moreover, when applying STBC to the periodic C-preamble depicted in Fig. 1, the periodicity of the C-preamble gets lost. This is due to the fact that the one or more STBC data words relating to the second C64 OFDM symbol will no longer be equal to the one or more STBC data words relating to the first C64 OFDM symbol. The loss of periodicity, however, leads to the problem that the symbol synchronization algorithms which make use of a periodic structure within the preamble can no longer be employed. Also, the C32 OFDM symbol cannot serve any longer as a guard interval separating the OFDM symbols within the preamble. The reason therefore is that in case of multipath propagation the first C64 OFDM symbol interferes with the second C64 OFDM symbol which is no longer equal to the first C64 OFDM symbol.

The above problems and further problems not explicitly discussed above do not occur when the data words are multiplexed in accordance with the invention. In Fig. 2, the physical layer of a transceiver 10 which is adapted to implement the method according to the invention is illustrated. The transceiver 10 comprises a scrambler 12, an FEC coding unit 14, an interleaving unit 16, a mapping unit 18, an OFDM unit 20, a burst forming unit 22, a block code encoder 24, a multiplexer 26, a radio transmitter 30 and a control unit 32. The block

code encoder 24 and the multiplexer 26 together form an encoder/multiplexer unit 28.

The transceiver 10 depicted in Fig. 1 receives as input signal a PDU train from a data link control (DLC). Each PDU train consists of information bits which are to be framed into a physical burst, i.e. a sequence of OFDM symbols to be encoded, multiplexed and transmitted.

Upon receipt of a PDU train the transmission bit rate within the transceiver 10 is configured by choosing an appropriate physical mode based on a link adaption mechanism. A physical mode is characterized by a specific modulation scheme and a specific code rate. In the HIPERLAN/2 standard several different coherent modulation schemes like BPSK, QPSK, 16-QAM and optional 64-QAM are specified. Also, for forward error control, convolutional codes with code rates of 1/2, 9/16 and 3/4 are specified which are obtained by puncturing of a convolutional mother code of rate 1/2. The possible resulting physical modes are depicted in Fig. 3. The data rate ranging from 6 to 54 Mbit/s can be varied by using various signal alphabets for modulating the OFDM subcarriers and by applying different puncturing patterns to a mother convolutional code.

Once an appropriate physical mode has been chosen, the $N_{BPDU}$ information bits contained within the PDU train are scrambled with the length-127 scrambler 12. The scrambled bits are then output to the FEC coding unit 14 which encodes the $N_{BPDU}$ scrambled PDU bits according to the previously set forward error correction.

The encoded bits output by the FEC coding unit 14 are input into the interleaving unit 16 which interleaves the encoded bits by using the appropriate interleaving scheme for the selected physical mode. The interleaved bits are input into the mapping unit 18 where sub-carrier modulation by mapping the interleaved bits into modulation constellation points in ac-

cordance with the chosen physical mode is performed. As mentioned above, the OFDM subcarriers are modulated by using BPSK, QPSK, 16-QAM or 64-QAM modulation depending on the physical mode selected for data transmission.

The mapping unit 18 outputs a stream of complex valued subcarrier modulation symbols which are divided in the OFDM unit in groups of 48 complex numbers. In the OFDM unit a complex base band signal is produced by OFDM modulation as described in ETSI TS 101 475, Broadband Radio Access Networks (BRAN); HIPERLAN Type 2; Physical (PHY) Layer, V1.1.1 (2000-04).

The complex base band OFDM symbols generated within the OFDM unit 20, where pilot subcarriers are inserted, are input into the physical burst unit 22, where an appropriate preamble is appended to the PDU train and the physical burst is built. The physical burst produced by the physical burst unit 22 has a format as depicted in Fig. 1. The physical burst unit 22 thus outputs a sequence of complex base band OFDM symbols in the form of the physical burst to the block code encoder 24.

The function of the block code encoder 24 is now generally described with reference to Fig. 4. In general, the block code encoder 24 receives an input signal in the form of a sequence of vectors $\mathbf{X} = [X_1 X_2 \ldots X_K]^T$ of the length K. The block code encoder 24 encodes each vector $\mathbf{X}$ and outputs for each vector $\mathbf{X}$ a data block comprising a plurality of signal vectors $\mathbf{C}^{(1)}, \mathbf{C}^{(2)} \ldots, \mathbf{C}^{(M)}$ as depicted in Fig. 4. Each signal vector $\mathbf{C}^{(1)}$, $\mathbf{C}^{(2)} \ldots$, $\mathbf{C}^{(M)}$ corresponds to a single data word. Thus, the data block generated from the vector $\mathbf{X}$ comprises M data words wherein M is the number of transmitter antennas.

Each data word $\mathbf{C}^{(i)}$ with $i = 1 \ldots M$ comprises N data symbols, i.e. each data word $\mathbf{C}^{(i)}$ has a length of N. The value of N cannot be freely chosen since the matrix $\mathbf{C}$ spanned by the data words $\mathbf{C}^{(i)}$ has to be orthogonal in this embodiment. Several examples for data blocks in the form of orthogonal code

matrices **C** are described in US 6,088,408. In the block coding approach described in the present embodiment all data symbols $c_j^i$ of the code matrix **C** are derived from the components of the input vector **X** and are simple linear functions thereof or of its complex conjugate.

If a receive signal vector **Y** at one receive antenna is denoted by $\mathbf{Y} = [Y_1 Y_2 \ldots Y_N]^T$, the relationship between **Y** and the code matrix **C** is as follows:

$$\begin{bmatrix} Y_2 \\ \ldots \\ Y_N \end{bmatrix} = \begin{bmatrix} c_2 & \ldots & & c_2 \\ \ldots & & \ldots & \ldots \\ c_N^{(1)} & c_N^{(2)} & \ldots & c_N^{(M)} \end{bmatrix} \cdot \begin{bmatrix} h \\ \ldots \\ h^{(M)} \end{bmatrix} \qquad (5)$$

where $h^{(i)}$ represents the channel coefficient of the channel from the i-th transmit antenna to the receive antenna. A generalization to more receive antennas is straightforward.

In the following examples of possible block code matrices for two and three transmitter antennas, respectively, are discussed in more detail. The configuration of a wireless communication system with two transmit antennas and one receive antenna is depicted in Fig. 5. For two transmit antennas one possible block code matrix **C** with a code rate R = 1 is :

$$\mathbf{C} = \begin{bmatrix} X_1 & X_2 \\ -X_2^* & X_2^* \end{bmatrix} \qquad (6)$$

For three transmit antennas one possible block code matrix **C** with a code rate R = 0,5 is:

$$\mathbf{C} = \begin{bmatrix} -X_4 & -X_3 & X_2 \\ X_1^* & X_2^* & X_3^* \\ -X_2^* & X_1^* & -X_4^* \\ -X_3^* & X_4^* & X_1^* \\ -X_4^* & -X_3^* & X_2^* \end{bmatrix} \qquad (7)$$

The code rate R is defined as the ratio of the length K of the input vector **X** and the length N of each code word $\mathbf{C}^{(i)}$:

$$R = K/N \qquad (8)$$

As can be seen from Fig. 4, the block code encoder 24 outputs for each data signal in the form of a vector **X** a data block in the form of a matrix **C**. The data block output by the block code encoder 24 is input into the multiplexer 26 which multiplexes the data words (vectors $\mathbf{C}^{(i)}$) of each data block in accordance with an externally provided control signal either in the time domain or in the frequency domain. The control signal is generated by the control unit 32 based on an assessment of the transmission constraints. The assessment of the transmission constraints and the controlling of the multiplexer 26 by means of the control unit 32 will be described later in more detail.

In the multicarrier scheme OFDM, the output of the block code encoder 24 is modulated onto subcarriers which are orthogonal to each other. There exist essentially two possibilities to

multiplex a data block comprising individual data words in an OFDM system. According to a first possibility depicted in Fig. 6, the data words of a specific data block are extended in the time direction (STBC). In other words: The data words are multiplexed in the time domain. According to a second possibility, the data words of a data block are extended in the frequency direction as depicted in Fig. 7. This means that the data words are multiplexed in the frequency domain. Multiplexing the data words of a data block in the form of a code matrix in the frequency domain will in the following be referred to as space-frequency block coding (SFBC).

As can be seen from Figs. 6 and 7, the individual data words of a data block are transmitted from different transmit antennas. According to the multiplexing scheme of Fig. 6, an individual data block is transmitted on an individual subcarrier over a time interval of $N \cdot T$, wherein N is the number of data symbols per data word and T is the duration of one of the data symbols. According to the multiplexing scheme of Fig. 7, an individual data block is spread over N subcarriers and is transmitted during a time interval of T. It can clearly be seen that the multiplexing scheme of Fig. 6 can generally be employed when the relation (4) is fulfilled and the multiplexing scheme of Fig. 7 can generally be employed when the relation (3) is satisfied.

The encoded and multiplexed output signal of the encoder/multiplexer unit 28 is input into the radio transmitter 30. The radio transmitter 30 performs radio transmission over a plurality of transmit antennas by modulating a radio frequency carrier with the output signal of the encoder/multiplexer unit 28. The transceiver 10 of Fig. 2 further comprises a receiver stage not depicted in Fig. 2. The receiver stage has a physical layer with components for performing the inverse operations of the components depicted in Fig. 2. For example, the receiver stage comprises a descram-

bler, a FEC decoding unit, a demultiplexer/decoder unit with a demultiplexer and a block code decoder, etc.

Now, the control of the multiplexer 26 will be described in more detail with reference to both physical and data-related transmission constraints that may occur if physical bursts as the one depicted in Fig. 1 are employed. In accordance with typical HIPERLAN/2 scenarios, it is supposed that relation (4) is fulfilled and that it cannot always be guaranteed that relation (3) is fulfilled. This corresponds to the realistic situation that the basic performance of STBC transmission is better than the basic performance of SFBC transmission. Basic performance here means that only physical transmission constraints are taken into account. In such a case the control unit 32 may decide that the data blocks have to be multiplexed in the time domain. However, if the physical transmission parameters change, there might occur the case where relation (4) is no longer fulfilled whereas relation (3) is fulfilled at least approximately. In this case the control unit 32 will decide that the data words of the data blocks are no longer multiplexed in the time domain. Instead, the control unit 32 controls the multiplexer 26 such that the data words of the data blocks are multiplexed in the frequency domain.

So far only physical transmission constraints have been considered. Should data-related transmission constraints also be of importance, the control unit 32 controls the multiplexer 26 by additionally taking into account data-related transmission constraints.

It has been mentioned above that the transmission constraints which have to be considered in context with the physical burst depicted in Fig. 1 are the preservation of a periodic structure in the C-preamble and the provision of a predefined number N of data symbols in each data word which is to be

multiplexed in the time domain. These two data-related transmission constraints can occur in several combinations.

According to a first scenario, the data signal has the structure of the physical burst depicted in Fig. 1 and comprises a user data section and a preamble with a periodic structure. It is further supposed that the data-related transmission constraint of preserving the periodic structure has to be taken into account while no data symbol fitting problem occurs with respect to the user data section. In such a case the data words of data blocks relating to the preamble are multiplexed in accordance with SFBC in the frequency domain and the data words of data blocks relating to the user data section are multiplexed in accordance with STBC in the time domain. By multiplexing the data words derived from the preamble in the frequency domain a preservation of the order of the C32 OFDM symbols and the two C64 OFDM symbols can be achieved.

According to a second scenario derived from the physical burst depicted in Fig. 1, the periodic structure within the preamble has to be preserved and additionally the data symbol fitting problem has to be taken into account with respect to the user data section. Like in the first scenario, the data words of data blocks derived from the preamble are multiplexed in accordance with SFBC in the frequency domain. Due to the data symbol fitting problem the data words of the last data block relating to the user data structure contains less than the predefined number N of data symbols contained in data words of the previous data blocks. Therefore, only the data words (containing the predefined number N of data symbols) of the previous data blocks are multiplexed in accordance with STBC in the time domain. The data words of the last data block contain $N_R = \mathrm{mod}(N_D/N)$ data symbols and are multiplexed in accordance with SFBC in the frequency domain, wherein $N_D$ is the total number of data symbols to be transmitted over one transmit antenna.

According to a third scenario, the data-related transmission constraint of the preservation of a periodic structure within the preamble is not relevant but the data symbol fitting problem is relevant with respect to the user data section. In this case the data words of data blocks relating to the preamble are multiplexed in accordance with STBC in the time domain and the data words of data blocks relating to the user data section are multiplexed as described above for the second scenario. In other words: The data words of the last data block have a length of $N_R$ data symbols and the data words of the previous data blocks have the predefined length of N data symbols.

According to a fourth scenario, the data-related transmission constraint of preserving a periodic structure has not to be taken into account and the physical transmission constraint of $B_C \gg N/T$ is at least approximately fulfilled. In this case the data words of data blocks relating to the preamble are multiplexed in accordance with STBC in the time domain and the data words of data blocks relating to the user data section are multiplexed in accordance with SFBC in the frequency domain. By using STBC for the preamble a good channel estimation can be performed. Due to the use of STBC for the preamble the slightly worse performance of SFBC can be compensated by means of receiver algorithms for interference suppression based on the good channel estimation. Using STBC for the preamble and SFBC for the user data section has the advantage that data symbol fitting problems with respect to the user data section do not appear.

Additional scenarios based on further combinations of data-related and physical transmission constraints can easily be realized in accordance with the invention. Also, the invention can easily be applied to data signals having a structure different from the structure of the physical burst depicted in Fig. 1. Although the invention is preferably practiced

with the transmit diversity scheme of a combination of STBC and SFBC, other transmit diversity schemes can be used as well.

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CLAIMS

1. A method of multiplexing data words in a multicarrier transmit diversity system, comprising:
   a) generating a plurality of data blocks (**C**), each data block (**C**) comprising data words ($\mathbf{C}^{(i)}$) and each data word ($\mathbf{C}^{(i)}$) containing data symbols ($c_j^{(i)}$) derived from a data signal;
   b) determining for one or more data blocks (**C**) in dependence on at least one transmission constraint if the data words ($\mathbf{C}^{(i)}$) of said one or more data blocks (**C**) are to be multiplexed in the time domain or in the frequency domain; and
   c) multiplexing the data words ($\mathbf{C}^{(i)}$) of the data blocks (**C**) in accordance with the determination in step b).

2. The method according to claim 1, wherein the data signal comprises at least one of a preamble and a user data section.

3. The method according to claim 1 or 2, wherein the at least one transmission constraint comprises a data-related transmission constraint.

4. The method according to claim 3, wherein the data-related transmission constraint is a pre-defined number (N) of data symbols ($c_j^{(i)}$) to be comprised within each data word ($\mathbf{C}^{(i)}$) which is to be multiplexed in the time domain.

5. The method according to claim 4, wherein the data words ($\mathbf{C}^{(i)}$) containing the predefined number (N) of data symbols ($c_j^{(i)}$) are multiplexed in the time domain and the data words ($\mathbf{C}^{(i)}$) containing more or

less data symbols ($c_j^{(i)}$) are multiplexed in the frequency domain.

6. The method according to claim 4 or 5, wherein the data signal or a portion thereof has a predefined length and wherein integer multiples of the predefined number of data symbols ($c_j^{(i)}$) are arranged in data blocks (**C**) with data words ($\mathbf{C}^{(i)}$) which are multiplexed in the time domain and a remainder of data symbols ($c_j^{(i)}$) is arranged in data blocks (**C**) with data words ($\mathbf{C}^{(i)}$) which are multiplexed in the frequency domain.

7. The method according to claim 6, wherein the user data section of the data signal has the predefined length.

8. The method according to claim 7, wherein the data words ($\mathbf{C}^{(i)}$) of data blocks (**C**) relating to the preamble are either multiplexed completely in the frequency domain or completely in the time domain dependent on the transmission constraint.

9. The method according to one of claims 1 to 8, wherein the data signal comprises one or more periodic structures (C32, C64).

10. The method according to claim 9, wherein the one or more periodic structures (C32, C64) are contained within the preamble.

11. The method according to claim 9 or 10, wherein the data-related transmission constraint is a preservation of the one or more periodic structures (C32, C64).

12. The method according to one of claims 9 to 11,

wherein at least the data words ($\mathbf{C}^{(i)}$) of data blocks ($\mathbf{C}$) relating to the periodic structures (C32, C64) are multiplexed in the frequency domain.

13. The method according to claim 12, wherein the data words ($\mathbf{C}^{(i)}$) of data blocks ($\mathbf{C}$) relating to the user data section are multiplexed in the time domain.

14. The method according to one of claims 1 to 13, wherein the at least one transmission constraint comprises a physical transmission constraint.

15. The method according to claim 14, wherein the physical transmission constraint is determined based on at least one of a coherence bandwidth and a coherence time.

16. The method according to claim 15, wherein the physical transmission constraint is determined by assessing if the relationship $B_C >> N/T$ is fulfilled, wherein $B_c$ is the coherence bandwidth, N is the number of data symbols ($c_j^{(i)}$) per data word ($\mathbf{C}^{(i)}$) and T is the duration of one of the data symbols ($c_j^{(i)}$).

17. The method according to claim 15 or 16, wherein the physical transmission constraint is determined by assessing if the relationship $t_C >> N \cdot T$ is fulfilled, wherein $t_c$ is the coherence time, N is the number of data symbols ($c_j^{(i)}$) per data word ($\mathbf{C}^{(i)}$) and T is the duration of one of the data symbols ($c_j^{(i)}$).

18. The method according to claim 16 or 17, wherein, when the physical transmission constraint $B_C >> N/T$ is at least approximately fulfilled, the data words ($\mathbf{C}^{(i)}$) of data blocks ($\mathbf{C}$) relating to the preamble are multiplexed in the time domain and the data words

($\mathbf{C}^{(i)}$) of data blocks ($\mathbf{C}$) relating to the user data sequence are multiplexed in the frequency domain.

19. The method according to one of claims 1 to 18, wherein the data blocks ($\mathbf{C}$) are obtained from the data signal by means of block coding or by means of permutation.

20. The method according to one of claims 1 to 19, wherein the data symbols ($c_j^{(i)}$) are modulated onto subcarriers which are orthogonal to each other.

21. A multiplexer (26) adapted to multiplex data words in accordance with the method according to one of claims 1 to 20.

22. A demultiplexer adapted to demultiplex data words multiplexed by the multiplexer of claim 21.

23. A transceiver for wireless communication, comprising at least one of a multiplexer according to claim 21 and a demultiplexer according to claim 22.

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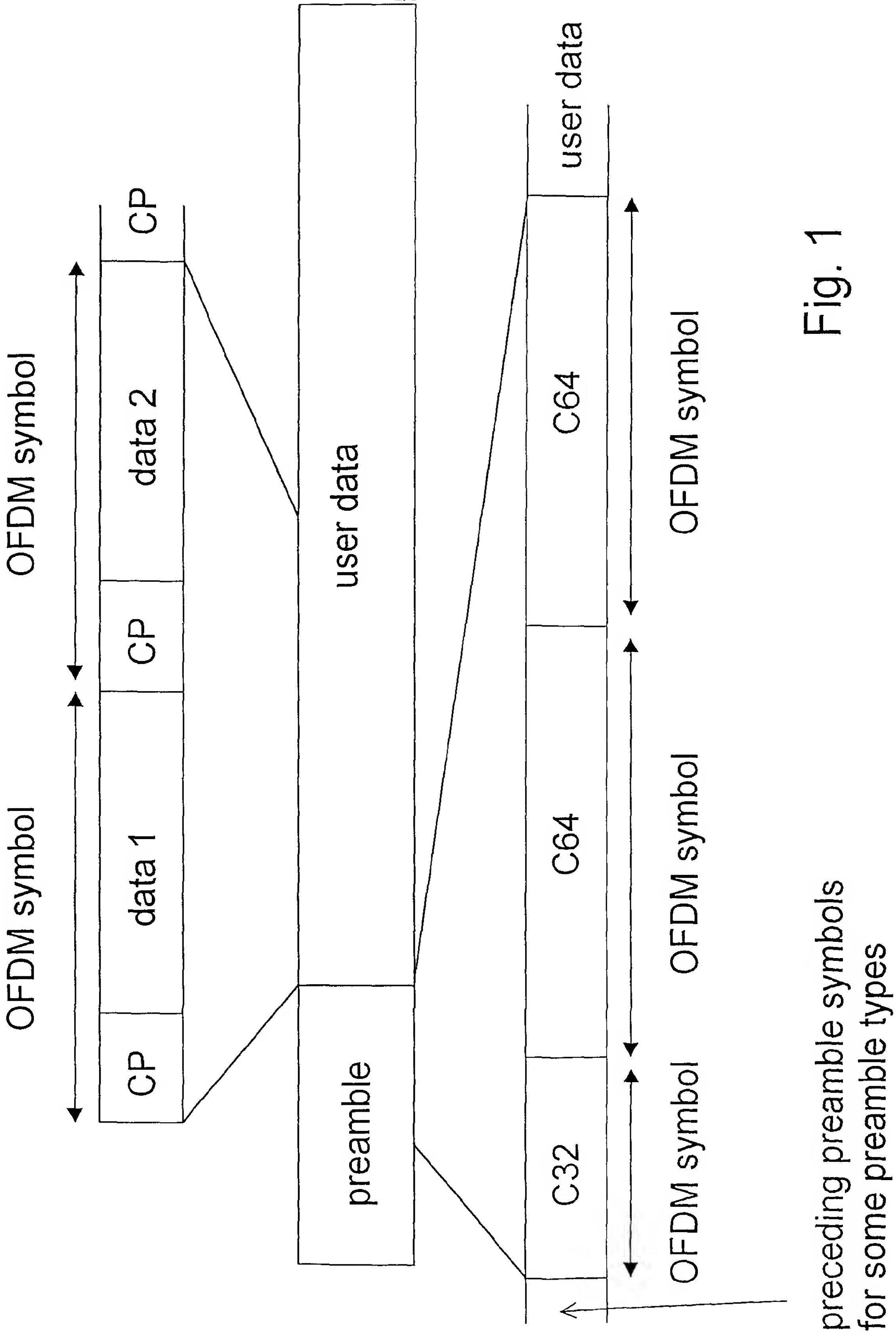
OFDM symbol
OFDM symbol
CP
data 1
CP
data 2
CP
preamble
user data
C32
C64
C64
user data
OFDM symbol
OFDM symbol
OFDM symbol
preceding preamble symbols
for some preamble types

Fig. 1

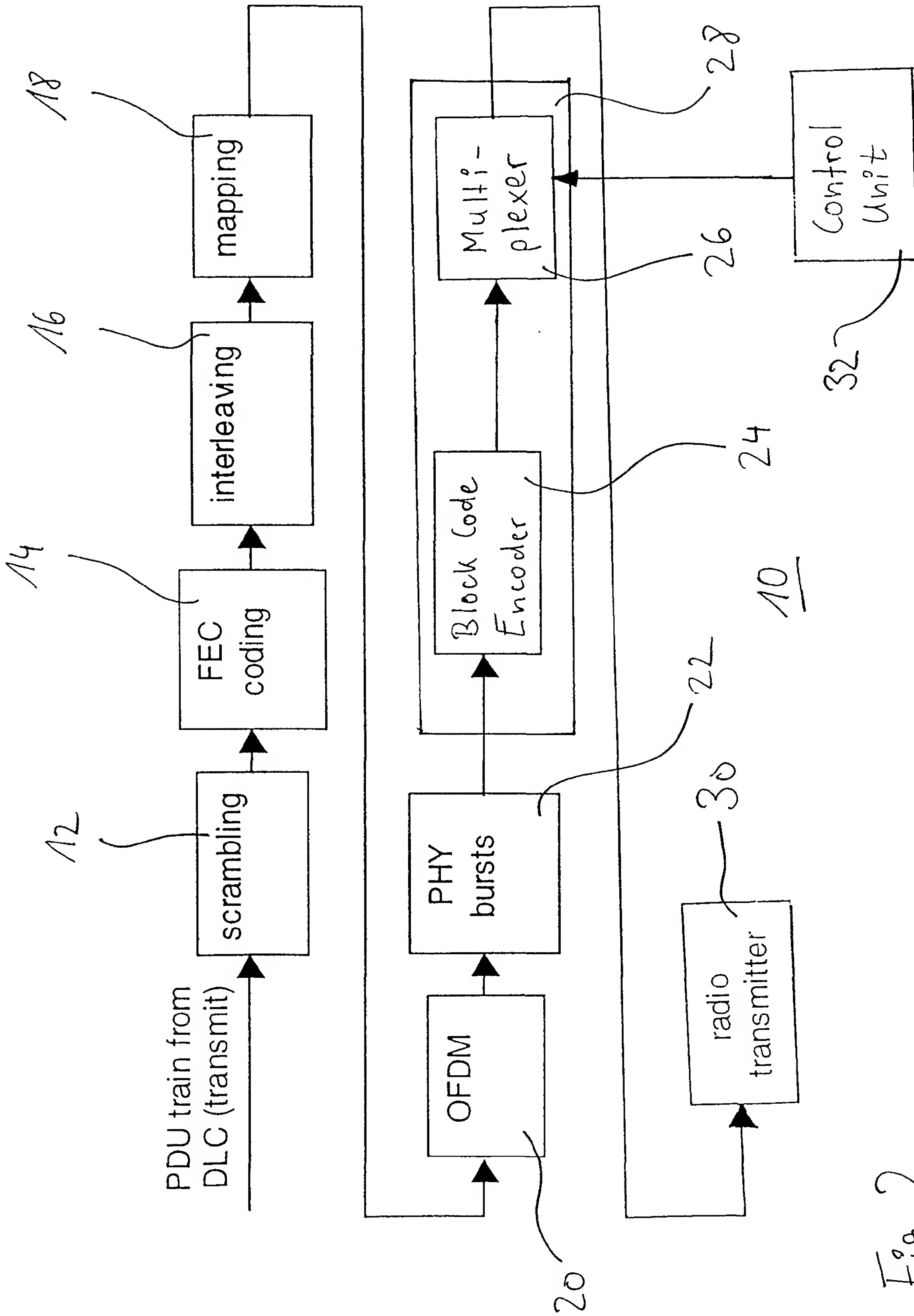
PDU train from DLC (transmit)
scrambling
FEC coding
interleaving
mapping
OFDM
PHY bursts
Block Code Encoder
Multi-plexer
Control Unit
radio transmitter
12
14
16
18
20
22
24
26
28
30
32
10

Fig. 2

| modulation scheme | code rate | bit rate |
|---|---|---|
| BPSK | 1/2 | 6 Mbps |
| BPSK | 3/4 | 9 Mbps |
| QPSK | 1/2 | 12 Mbps |
| QPSK | 3/4 | 18 Mbps |
| 16-QAM | 9/16 | 27 Mbps |
| 16-QAM | 3/4 | 36 Mbps |
| 64-QAM | 3/4 | 54 Mbps |

Fig. 3

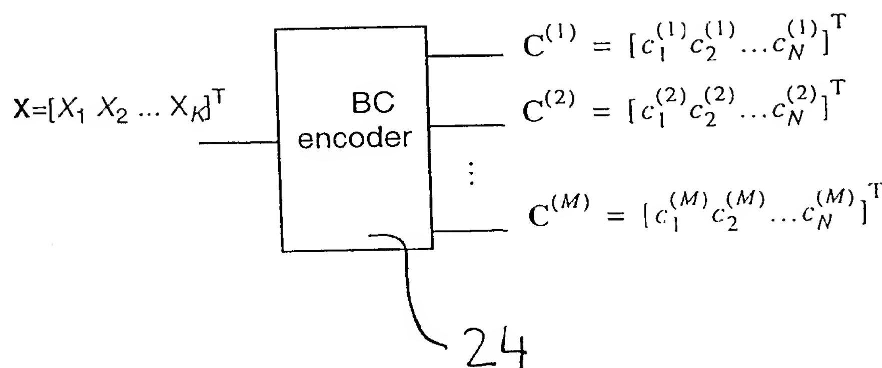


Fig. 4

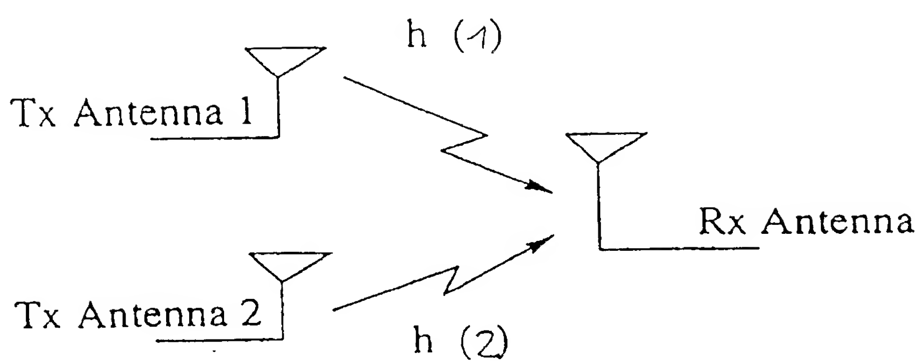


Fig. 5

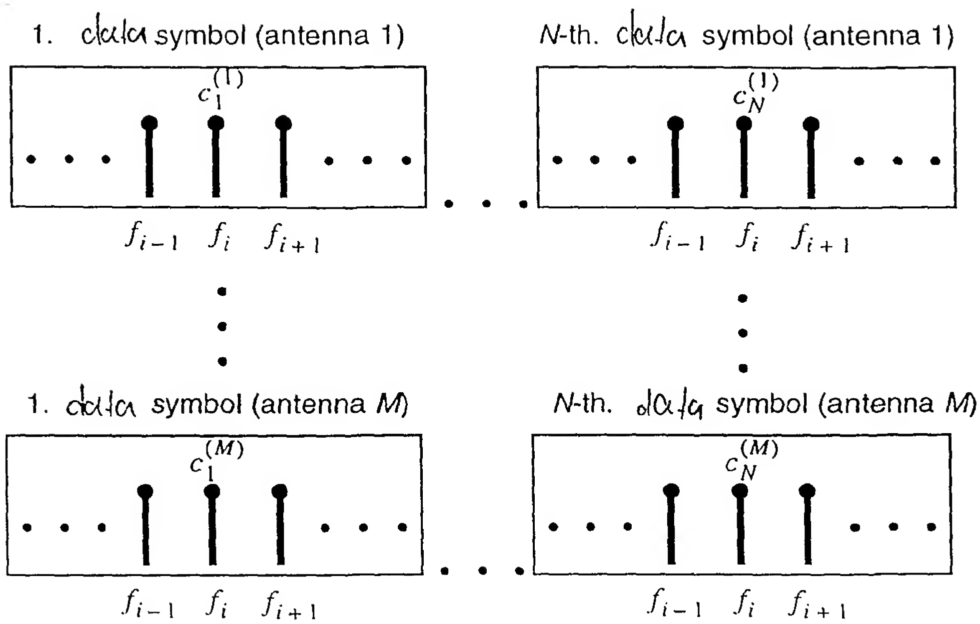
1. data symbol (antenna 1)
N-th. data symbol (antenna 1)
$c_1^{(1)}$
$c_N^{(1)}$
$f_{i-1}$ $f_i$ $f_{i+1}$
1. data symbol (antenna M)
N-th. data symbol (antenna M)
$c_1^{(M)}$
$c_N^{(M)}$

Fig. 6

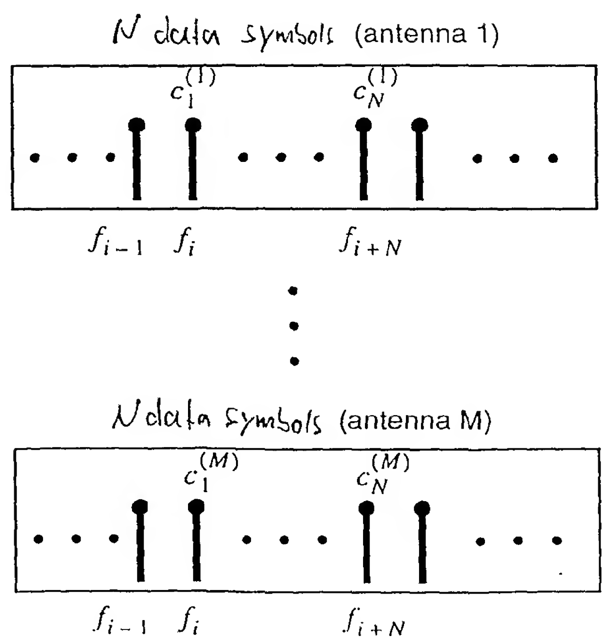
N data symbols (antenna 1)
$c_1^{(1)}$
$c_N^{(1)}$
$f_{i-1}$ $f_i$ $f_{i+N}$
N data symbols (antenna M)
$c_1^{(M)}$
$c_N^{(M)}$

Fig. 7

# INTERNATIONAL SEARCH REPORT

International Application No
PCT/EP 02/02245

**A. CLASSIFICATION OF SUBJECT MATTER**
IPC 7 H04L5/02 H04L1/00 H04B7/06 H04B7/12 H04L1/06
H04L1/04

According to International Patent Classification (IPC) or to both national classification and IPC

**B. FIELDS SEARCHED**

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, WPI Data, PAJ, INSPEC

**C. DOCUMENTS CONSIDERED TO BE RELEVANT**

| Category ° | Citation of document, with indication, where appropriate, of the relevant passages | Relevant to claim No. |
|---|---|---|
| X | MUDULODU S ET AL: "A transmit diversity scheme for frequency selective fading channels" GLOBECOM '00 - IEEE GLOBAL TELECOMMUNICATIONS CONFERENCE. CONFERENCE RECORD (CAT. NO.00CH37137), PROCEEDINGS OF GLOBAL TELECOMMUNICATIONS CONFERENCE, SAN FRANCISCO, CA, USA, 27 NOV.-1 DEC. 2000, pages 1089-1093 vol.2, XP002172508 2000, IEEE, Piscataway, NJ, USA ISBN: 0-7803-6451-1 cited in the application the whole document<br>---<br>-/-- | 1-23 |

[X] Further documents are listed in the continuation of box C.

[ ] Patent family members are listed in annex.

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"&" document member of the same patent family

| Date of the actual completion of the international search | Date of mailing of the international search report |
|---|---|
| 9 July 2002 | 17/07/2002 |

| Name and mailing address of the ISA | Authorized officer |
|---|---|
| European Patent Office, P.B. 5818 Patentlaan 2<br>NL – 2280 HV Rijswijk<br>Tel. (+31–70) 340–2040, Tx. 31 651 epo nl,<br>Fax: (+31–70) 340–3016 | Toumpoulidis, T |

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

| Category ° | Citation of document, with indication,where appropriate, of the relevant passages | Relevant to claim No. |
|---|---|---|
| X | LEE K F ET AL: "A space-frequency transmitter diversity technique for OFDM systems" GLOBECOM '00 - IEEE. GLOBAL TELECOMMUNICATIONS CONFERENCE. CONFERENCE RECORD (CAT. NO.00CH37137), PROCEEDINGS OF GLOBAL TELECOMMUNICATIONS CONFERENCE, SAN FRANCISCO, CA, USA, 27 NOV.-1 DEC. 2000, pages 1473-1477 vol.3, XP002172509 2000, Piscataway, NJ, USA, IEEE, USA ISBN: 0-7803-6451-1 the whole document<br>--- | 1-23 |
| A | ALAMOUTI S M: "A simple transmit diversity technique for wireless communications" IEEE JOURNAL ON SELECTED AREAS IN COMMUNICATIONS,IEEE INC. NEW YORK,US, vol. 16, no. 8, October 1998 (1998-10), pages 1451-1458, XP002100058 ISSN: 0733-8716 the whole document<br>----- | 1-23 |

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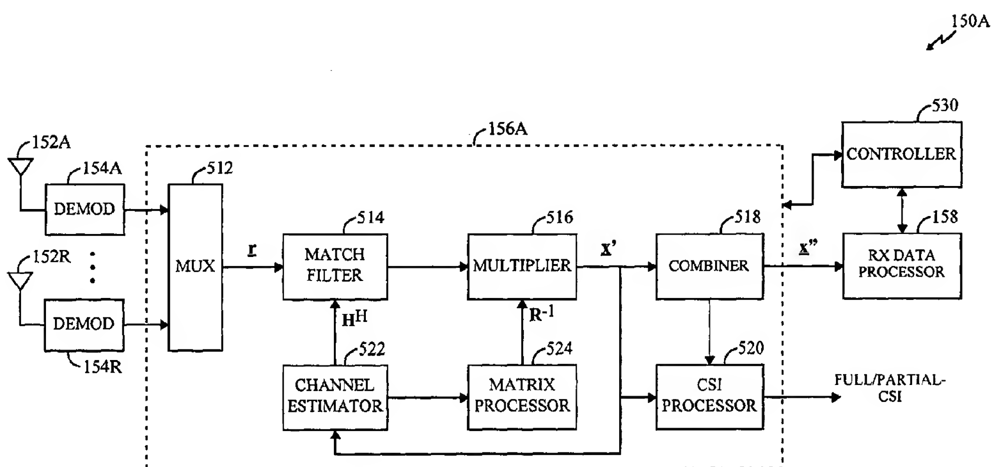
*[Continued on next page]*

(54) **Title:** METHOD AND APPARATUS FOR UTILIZING CHANNEL STATE INFORMATION IN A WIRELESS COMMUNICATION SYSTEM




(57) **Abstract:** Techniques for transmitting data from a transmitter unit to a receiver unit in a multiple-input multiple-output (MIMO) communication system. In one method, at the receiver unit, a number of signals are received via a number of receive antennas, with the received signal from the transmitter unit. The received signals are processed to derive channel state information (CSI) indicative of characteristics of a number of transmission channels used for data transmission. The CSI is transmitted back to the transmitter unit. At the transmitter unit, the CSI from the receiver unit is received and data for transmission to the receiver units is processed based on the received CSI.

WO 02/078211 A2

*For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.*

# METHOD AND APPARATUS FOR UTILIZING CHANNEL STATE INFORMATION IN A WIRELESS COMMUNICATION SYSTEM

## BACKGROUND

### Field

**[1001]** The present invention relates generally to data communication, and more specifically to a novel and improved method and apparatus for utilizing (full or partial) channel state information to provide improved performance for a wireless communication system.

### Background

**[1002]** Wireless communication systems are widely deployed to provide various types of communication such as voice, data, and so on. These systems may be based on code division multiple access (CDMA), time division multiple access (TDMA), orthogonal frequency division modulation (OFDM), or some other modulation techniques. OFDM systems can provide high performance for some channel environments.

**[1003]** In a terrestrial communication system (e.g., a cellular system, a broadcast system, a multi-channel multi-point distribution system (MMDS), and others), an RF modulated signal from a transmitter unit may reach a receiver unit via a number of transmission paths. The characteristics of the transmission paths typically vary over time due to a number of factors such as fading and multipath.

**[1004]** To provide diversity against deleterious path effects and improve performance, multiple transmit and receive antennas may be used. If the transmission paths between the transmit and receive antennas are linearly independent (i.e., a transmission on one path is not formed as a linear combination of the transmissions on other paths), which is generally true to some extent, then the likelihood of correctly receiving a transmitted signal increases as the number of antennas increases. Generally, diversity increases and performance improves as the number of transmit and receive antennas increases.

[1005] A multiple-input multiple-output (MIMO) communication system employs multiple ($N_T$) transmit antennas and multiple ($N_R$) receive antennas for data transmission. A MIMO channel may be decomposed into $N_C$ independent channels, with $N_C \leq \min \{N_T, N_R\}$. Each of the $N_C$ independent channels is also referred to as a spatial subchannel of the MIMO channel and corresponds to a dimension. The MIMO system can provide improved performance if the additional dimensionalities created by the multiple transmit and receive antennas are utilized.

[1006] There is therefore a need in the art for techniques to utilize channel state information (CSI) to take advantage of the additional dimensionalities created by a MIMO system to provide improved system performance.

## SUMMARY

[1007] Aspects of the invention provide techniques to process received signals in a multiple-input multiple-output (MIMO) communication system to recover transmitted signals, and to estimate the characteristics of a MIMO channel. Various receiver processing schemes may be used to derive channel state information (CSI) indicative of the characteristics of the transmission channels used for data transmission. The CSI is then reported back to the transmitter system and used to adjust the signal processing (e.g., coding, modulation, and so on). In this manner, high performance is achieved based on the determined channel conditions.

[1008] A specific embodiment of the invention provides a method for transmitting data from a transmitter unit to a receiver unit in a MIMO communication system. In accordance with the method, at the receiver unit, a number of signals are received via a number of receive antennas, with the received signal from each receive antenna comprising a combination of one or more signals transmitted from the transmitter unit. The received signals are processed (e.g., via a channel correlation matrix inversion (CCMI) scheme, an unbiased minimum mean square error (UMMSE) scheme, or some other receiver processing scheme) to derive CSI indicative of characteristics of a number of transmission channels used for data transmission. The CSI is encoded and transmitted back to the transmitter unit. At the transmitter unit, the CSI from the receiver unit is received and data for transmission to the receiver unit is processed based on the received CSI.

[1009] The reported CSI may include full CSI or partial CSI. Full CSI includes sufficient full-bandwidth characterization (e.g., the amplitude and phase across the

useable bandwidth) of the propagation path between all pairs of transmit and receive antennas. Partial CSI may include, for example, the signal-to-noise-plus-interference (SNR) of the transmission channels. At the transmitter unit, the data for each transmission channel may be coded based on the SNR estimate for the transmission channel, and the coded data for each transmission channel may be modulated in accordance with a modulation scheme selected based on the SNR estimate. For full-CSI processing, the modulation symbols are also pre-processed prior to transmission in accordance with the received CSI.

**[1010]** The invention further provides methods, systems, and apparatus that implement various aspects, embodiments, and features of the invention, as described in further detail below.

## BRIEF DESCRIPTION OF THE DRAWINGS

**[1011]** The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

**[1012]** FIG. 1 is a diagram of a multiple-input multiple-output (MIMO) communication system capable of implementing various aspects and embodiments of the invention;

**[1013]** FIGS. 2A and 2B are block diagrams of an embodiment of a MIMO transmitter system capable of performing partial-CSI processing and full-CSI processing, respectively;

**[1014]** FIG. 3 is a block diagram of an embodiment of a MIMO transmitter system which utilizes orthogonal frequency division modulation (OFDM);

**[1015]** FIG. 4 is a block diagram of a portion of a MIMO transmitter system capable of providing different processing for different transmission types and which also employs OFDM;

**[1016]** FIGS. 5 and 6 are block diagrams of two embodiments of a receiver system having multiple ($N_R$) receive antennas and capable of processing a data transmission based on a channel correlation matrix inversion (CCMI) technique and an unbiased minimum mean square error (UMMSE), respectively;

**[1017]** FIG. 7A shows the average throughput for the MIMO system for three receiver processing techniques and for different SNR values; and

[1018] FIG. 7B shows the cumulative probability distribution functions (CDF) for the three receiver processing techniques generated based on the histogram of the data.

## DETAILED DESCRIPTION

[1019] FIG. 1 is a diagram of a multiple-input multiple-output (MIMO) communication system 100 capable of implementing various aspects and embodiments of the invention. System 100 includes a first system 110 in communication with a second system 150. System 100 can be operated to employ a combination of antenna, frequency, and temporal diversity (described below) to increase spectral efficiency, improve performance, and enhance flexibility. In an aspect, system 150 can be operated to determine the characteristics of the communication link and to report channel state information (CSI) back to system 110, and system 110 can be operated to adjust the processing (e.g., encoding and modulation) of data to be transmitted based on the reported CSI.

[1020] Within system 110, a data source 112 provides data (i.e., information bits) to a transmit (TX) data processor 114, which encodes the data in accordance with a particular encoding scheme, interleaves (i.e., reorders) the encoded data based on a particular interleaving scheme, and maps the interleaved bits into modulation symbols for one or more transmission channels used for transmitting the data. The encoding increases the reliability of the data transmission. The interleaving provides time diversity for the coded bits, permits the data to be transmitted based on an average signal-to-noise-plus-interference (SNR) for the transmission channels used for the data transmission, combats fading, and further removes correlation between coded bits used to form each modulation symbol. The interleaving may further provide frequency diversity if the coded bits are transmitted over multiple frequency subchannels. In accordance with an aspect of the invention, the encoding, interleaving, and symbol mapping (or a combination thereof) are performed based on the full or partial CSI available to system 110, as indicated in FIG. 1.

[1021] The encoding, interleaving, and symbol mapping at transmitter system 110 can be performed based on numerous schemes. One specific scheme is described in U.S Patent Application Serial No. 09/776,073, entitled "CODING SCHEME FOR A WIRELESS COMMUNICATION SYSTEM," filed February 1, 2001, assigned to the assignee of the present application and incorporated herein by reference.

[1022] MIMO system 100 employs multiple antennas at both the transmit and receive ends of the communication link. These transmit and receive antennas may be used to provide various forms of spatial diversity, including transmit diversity and receive diversity. Spatial diversity is characterized by the use of multiple transmit antennas and one or more receive antennas. Transmit diversity is characterized by the transmission of data over multiple transmit antennas. Typically, additional processing is performed on the data transmitted from the transmit antennas to achieved the desired diversity. For example, the data transmitted from different transmit antennas may be delayed or reordered in time, coded and interleaved across the available transmit antennas, and so on. Receive diversity is characterized by the reception of the transmitted signals on multiple receive antennas, and diversity is achieved by simply receiving the signals via different signal paths.

[1023] System 100 may be operated in a number of different communication modes, with each communication mode employing antenna, frequency, or temporal diversity, or a combination thereof. The communication modes may include, for example, a "diversity" communication mode and a "MIMO" communication mode. The diversity communication mode employs diversity to improve the reliability of the communication link. In a common application of the diversity communication mode, which is also referred to as a "pure" diversity communication mode, data is transmitted from all available transmit antennas to a recipient receiver system. The pure diversity communications mode may be used in instances where the data rate requirements are low or when the SNR is low, or when both are true. The MIMO communication mode employs antenna diversity at both ends of the communication link (i.e., multiple transmit antennas and multiple receive antennas) and is generally used to both improve the reliability and increase the capacity of the communications link. The MIMO communication mode may further employ frequency and/or temporal diversity in combination with the antenna diversity.

[1024] System 100 may further utilize orthogonal frequency division modulation (OFDM), which effectively partitions the operating frequency band into a number of (L) frequency subchannels (i.e., frequency bins). At each time slot (i.e., a particular time interval that may be dependent on the bandwidth of the frequency subchannel), a modulation symbol may be transmitted on each of the L frequency subchannels.

[1025] System 100 may be operated to transmit data via a number of transmission channels. As noted above, a MIMO channel may be decomposed into $N_C$ independent

channels, with $N_C \leq \min\{N_T, N_R\}$. Each of the $N_C$ independent channels is also referred to as a spatial subchannel of the MIMO channel. For a MIMO system not utilizing OFDM, there may be only one frequency subchannel and each spatial subchannel may be referred to as a "transmission channel". For a MIMO system utilizing OFDM, each spatial subchannel of each frequency subchannel may be referred to as a transmission channel. And for an OFDM system not operated in the MIMO communication mode, there is only one spatial subchannel and each frequency subchannel may be referred to as a transmission channel.

**[1026]** A MIMO system can provide improved performance if the additional dimensionalities created by the multiple transmit and receive antennas are utilized. While this does not necessarily require knowledge of CSI at the transmitter, increased system efficiency and performance are possible when the transmitter is equipped with CSI, which is descriptive of the transmission characteristics from the transmit antennas to the receive antennas. CSI may be categorized as either "full CSI" or "partial CSI".

**[1027]** Full CSI includes sufficient characterization (e.g., the amplitude and phase) across the entire system bandwidth (i.e., each frequency subchannel) for the propagation path between each transmit-receive antenna pair in the $N_T x N_R$ MIMO matrix. Full-CSI processing implies that (1) the channel characterization is available at both the transmitter and receiver, (2) the transmitter computes eigenmodes for the MIMO channel (described below), determines modulation symbols to be transmitted on the eigenmodes, linearly preconditions (filters) the modulation symbols, and transmits the preconditioned modulation symbols, and (3) the receiver performs a complementary processing (e.g., spatial matched filter) of the linear transmit processing based on the channel characterization to compute the $N_C$ spatial matched filter coefficients needed for each transmission channel (i.e., each eigenmode). Full-CSI processing further entails processing the data (e.g., selecting the proper coding and modulation schemes) for each transmission channel based on the channel's eigenvalue (described below) to derive the modulation symbols.

**[1028]** Partial CSI may include, for example, the signal-to-noise-plus-interference (SNR) of the transmission channels (i.e., the SNR for each spatial subchannel for a MIMO system without OFDM, or the SNR for each frequency subchannel of each spatial subchannel for a MIMO system with OFDM). Partial-CSI processing may imply processing the data (e.g., selecting the proper coding and modulation schemes) for each transmission channel based on the channel's SNR.

[1029] Referring to FIG. 1, a TX MIMO processor 120 receives and processes the modulation symbols from TX data processor 114 to provide symbols suitable for transmission over the MIMO channel. The processing performed by TX MIMO processor 120 is dependent on whether full or partial CSI processing is employed, and is described in further detail below.

[1030] For full-CSI processing, TX MIMO processor 120 may demultiplex and precondition the modulation symbols. And for partial-CSI processing, TX MIMO processor 120 may simply demultiplex the modulation symbols. The full and partial-CSI MIMO processing is described in further detail below. For a MIMO system employing full-CSI processing but not OFDM, TX MIMO processor 120 provides a stream of preconditioned modulation symbols for each transmit antenna, one preconditioned modulation symbol per time slot. Each preconditioned modulation symbol is a linear (and weighted) combination of $N_C$ modulation symbols at a given time slot for the $N_C$ spatial subchannels, as described in further detail below. For a MIMO system employing full-CSI processing and OFDM, TX MIMO processor 120 provides a stream of preconditioned modulation symbol vectors for each transmit antenna, with each vector including L preconditioned modulation symbols for the L frequency subchannels for a given time slot. For a MIMO system employing partial-CSI processing but not OFDM, TX MIMO processor 120 provides a stream of modulation symbols for each transmit antenna, one modulation symbol per time slot. And for a MIMO system employing partial-CSI processing and OFDM, TX MIMO processor 120 provides a stream of modulation symbol vectors for each transmit antenna, with each vector including L modulation symbols for the L frequency subchannels for a given time slot. For all cases described above, each stream of (either unconditioned or preconditioned) modulation symbols or modulation symbol vectors is received and modulated by a respective modulator (MOD) 122, and transmitted via an associated antenna 124.

[1031] In the embodiment shown in FIG. 1, receiver system 150 includes a number of receive antennas 152 that receive the transmitted signals and provide the received signals to respective demodulators (DEMOD) 154. Each demodulator 154 performs processing complementary to that performed at modulator 122. The demodulated symbols from all demodulators 154 are provided to a receive (RX) MIMO processor 156 and processed in a manner described below. The received modulation symbols for the transmission channels are then provided to a RX data processor 158, which performs

processing complementary to that performed by TX data processor 114. In a specific design, RX data processor 158 provides bit values indicative of the received modulation symbols, deinterleaves the bit values, and decodes the deinterleaved values to generate decoded bits, which are then provided to a data sink 160. The received symbol de-mapping, deinterleaving, and decoding are complementary to the symbol mapping, interleaving, and encoding performed at transmitter system 110. The processing by receiver system 150 is described in further detail below.

**[1032]** The spatial subchannels of a MIMO system (or more generally, the transmission channels in a MIMO system with or without OFDM) typically experience different link conditions (e.g., different fading and multipath effects) and may achieve different SNR. Consequently, the capacity of the transmission channels may be different from channel to channel. This capacity may be quantified by the information bit rate (i.e., the number of information bits per modulation symbol) that may be transmitted on each transmission channel for a particular level of performance. Moreover, the link conditions typically vary with time. As a result, the supported information bit rates for the transmission channels also vary with time. To more fully utilize the capacity of the transmission channels, CSI descriptive of the link conditions may be determined (typically at the receiver unit) and provided to the transmitter unit so that the processing can be adjusted (or adapted) accordingly. Aspects of the invention provide techniques to determine and utilize (full or partial) CSI to provide improved system performance.

## MIMO Transmitter System with Partial-CSI Processing

[1033] FIG. 2A is a block diagram of an embodiment of a MIMO transmitter system 110a, which is one embodiment of the transmitter portion of system 110 in FIG. 1. Transmitter system 110a (which does not utilize OFDM) is capable of adjusting its processing based on partial CSI reported by receiver system 150. System 110a includes (1) a TX data processor 114a that receives and processes information bits to provide modulation symbols and (2) a TX MIMO processor 120a that demultiplexes the modulation symbols for the $N_T$ transmit antennas.

[1034] TX data processor 114a is one embodiment of TX data processor 114 in FIG. 1, and many other designs may also be used for TX data processor 114 and are within the scope of the invention. In the specific embodiment shown in FIG. 2A, TX data processor 114a includes an encoder 202, a channel interleaver 204, a puncturer 206, and a symbol mapping element 208. Encoder 202 receives and encodes the information bits in accordance with a particular encoding scheme to provide coded bits. Channel interleaver 204 interleaves the coded bits based on a particular interleaving scheme to provide diversity. Puncturer 206 punctures zero or more of the interleaved coded bits to provide the desired number of coded bits. And symbol mapping element 208 maps the unpunctured coded bit into modulation symbols for one or more transmission channels used for transmitting the data.

[1035] Although not shown in FIG. 2A for simplicity, pilot data (e.g., data of known pattern) may be encoded and multiplexed with the processed information bits. The processed pilot data may be transmitted (e.g., in a time division multiplexed manner) in all or a subset of the transmission channels used to transmit the information bits. The pilot data may be used at the receiver to perform channel estimation, as is known in the art and described in further detail below.

[1036] As shown in FIG. 2A, the encoding and modulation may be adjusted based on the partial-CSI reported by receiver system 150. In an embodiment, adaptive encoding is achieved by using a fixed base code (e.g., a rate 1/3 Turbo code) and adjusting the puncturing to achieve the desired code rate, as supported by the SNR of the transmission channel used to transmit data. Alternatively, different coding schemes may be used based on the reported partial-CSI (as indicated by the dashed arrow into block 202). For example, each of the transmission channels may be coded with an independent code. With this coding scheme, a successive "nulling/equalization and interference cancellation" receiver processing scheme may be used to detect and decode

the data streams to derive a more reliable estimate of the transmitted data streams. One such receiver processing scheme is described by P.W. Wolniansky, et al in a paper entitled "V-BLAST: An Architecture for Achieving Very High Data Rates over the Rich-Scattering Wireless Channel", Proc. ISSSE-98, Pisa, Italy, and incorporated herein by reference.

**[1037]** For each transmission channel, symbol mapping element 208 can be designed to group sets of unpunctured coded bits to form non-binary symbols, and to map the non-binary symbols into points in a signal constellation corresponding to a particular modulation scheme (e.g., QPSK, M-PSK, M-QAM, or some other scheme) selected for that transmission channel. Each mapped point corresponds to a modulation symbol. The number of information bits that may be transmitted for each modulation symbol for a particular level of performance (e.g., one percent packet error rate) is dependent on the SNR of the transmission channel. Thus, the coding scheme and modulation scheme for each transmission channel may be selected based on the reported partial-CSI. The channel interleaving may also be adjusted based on the reported partial-CSI (as indicated by the dashed arrow into block 204).

**[1038]** Table 1 lists various combinations of coding rate and modulation scheme that may be used for a number of SNR ranges. The supported bit rate for each transmission channel may be achieved using any one of a number of possible combinations of coding rate and modulation scheme. For example, one information bit per symbol may be achieved using (1) a coding rate of 1/2 and QPSK modulation, (2) a coding rate of 1/3 and 8-PSK modulation, (3) a coding rate of 1/4 and 16-QAM, or some other combination of coding rate and modulation scheme. In Table 1, QPSK, 16-QAM, and 64-QAM are used for the listed SNR ranges. Other modulation schemes such as 8-PSK, 32-QAM, 128-QAM, and so on, may also be used and are within the scope of the invention.

Table 1

| SNR Range | # of Information Bits/Symbol | Modulation Symbol | # of Coded Bits/Symbol | Coding Rate |
|---|---|---|---|---|
| 1.5 – 4.4 | 1 | QPSK | 2 | 1/2 |
| 4.4 – 6.4 | 1.5 | QPSK | 2 | 3/4 |
| 6.4 – 8.35 | 2 | 16-QAM | 4 | 1/2 |
| 8.35 – 10.4 | 2.5 | 16-QAM | 4 | 5/8 |

| | | | | |
|---|---|---|---|---|
| 10.4 – 12.3 | 3 | 16-QAM | 4 | 3/4 |
| 12.3 – 14.15 | 3.5 | 64-QAM | 6 | 7/12 |
| 14.15 – 15.55 | 4 | 64-QAM | 6 | 2/3 |
| 15.55 – 17.35 | 4.5 | 64-QAM | 6 | 3/4 |
| > 17.35 | 5 | 64-QAM | 6 | 5/6 |

**[1039]** The modulation symbols from TX data processor 114a are provided to a TX MIMO processor 120a, which is one embodiment of TX MIMO processor 120 in FIG. 1. Within TX MIMO processor 120a, a demultiplexer 214 demultiplexes the received modulation symbols into a number of ($N_T$) streams of modulation symbols, one stream for each antenna used to transmit the modulation symbols. Each stream of modulation symbols is provided to a respective modulator 122. Each modulator 122 converts the modulation symbols into an analog signal, and further amplifies, filters, quadrature modulates, and upconverts the signal to generate a modulated signal suitable for transmission over the wireless link.

**[1040]** If the number of spatial subchannels is less than the number of available transmit antennas (i.e., $N_C < N_T$) then various schemes may be used for the data transmission. In one scheme, $N_C$ modulation symbol steams are generated and transmitted on a subset (i.e., $N_C$) of the available transmitted antennas. The remaining ($N_T$ - $N_C$) transmit antennas are not used for the data transmission. In another scheme, the additional degrees of freedom provided by the ($N_T$ - $N_C$) additional transmit antennas are used to improve the reliability of the data transmission. For this scheme, each of one or more data streams may be encoded, possibly interleaved, and transmitted over multiple transmit antennas. The use of multiple transmit antennas for a data stream increases diversity and improves reliability against deleterious path effects.

## <u>MIMO Transmitter System with Full-CSI Processing</u>

**[1041]** FIG. 2B is a block diagram of an embodiment of a MIMO transmitter system 110b (which does not utilize OFDM) capable of processing data based on full CSI reported by receiver system 150. The information bits are encoded, interleaved, and symbol mapped by a TX data processor 114 to generate modulation symbols. The coding and modulation may be adjusted based on the available full-CSI reported by the receiver system, and may be performed as described above for MIMO transmitter system 110a.

**[1042]** Within a TX MIMO processor 120b, a channel MIMO processor 212 demultiplexes the received modulation symbols into a number of ($N_c$) modulation symbol streams, one stream for each spatial subchannel (i.e., eigenmode) used to transmit the modulation symbols. For full-CSI processing, channel MIMO processor 212 preconditions the $N_c$ modulation symbols at each time slot to generate $N_T$ preconditioned modulation symbols, as follows:

$$\begin{bmatrix} x_1 \\ x_2 \\ \mathrm{M} \\ x_{N_T} \end{bmatrix} = \begin{bmatrix} e_{11}, & e_{12}, & & e_{1N_c} \\ e_{21}, & e_{22}, & & e_{2N_c} \\ & & & \\ e_{N_T1}, & e_{N_T1}, & & e_{N_TN_c} \end{bmatrix} \cdot \begin{bmatrix} b_1 \\ b_2 \\ \mathrm{M} \\ b_{N_C} \end{bmatrix} \qquad \text{Eq (1)}$$

where $b_1$, $b_2$, ... and $b_{N_c}$ are respectively the modulation symbols for the spatial subchannels 1, 2, ... $N_{Nc}$, where each of the $N_C$ modulation symbols may be generated using, for example, M-PSK, M-QAM, or some other modulation scheme;

$e_{ij}$ are elements of an eigenvector matrix **E** related to the transmission characteristics from the transmit antennas to the receive antennas; and

$x_1$, $x_2$, ... $x_{N_T}$ are the preconditioned modulation symbols, which can be expressed as:

$$x_1 = b_1 \cdot e_{11} + b_2 \cdot e_{12} + \ \ldots\ + b_{N_C} \cdot e_{1N_C} \ ,$$

$$x_2 = b_1 \cdot e_{21} + b_2 \cdot e_{22} + \ \ldots\ + b_{N_C} \cdot e_{2N_C} \ , \text{ and}$$

$$x_{N_T} = b_1 \cdot e_{N_T1} + b_2 \cdot e_{N_T2} + \ \ldots\ + b_{N_C} \cdot e_{N_TN_C} \ .$$

The eigenvector matrix **E** may be computed by the transmitter or is provided to the transmitter by the receiver.

**[1043]** For full-CSI processing, each preconditioned modulation symbol, $x_i$, for a particular transmit antenna represents a linear combination of (weighted) modulation symbols for up to $N_C$ spatial subchannels. The modulation scheme employed for each of the modulation symbol $x_i$ is based on the effective SNR of that eigenmode and is proportional to an eigenvalue, $\lambda_i$ (described below). Each of the $N_C$ modulation symbols used to generate each preconditioned modulation symbol may be associated with a different signal constellation. For each time slot, the $N_T$ preconditioned modulation symbols generated by channel MIMO processor 212 are demultiplexed by a demultiplexer 214 and provided to $N_T$ modulators 122.

[1044] The full-CSI processing may be performed based on the available CSI and on the selected transmit antennas. The full-CSI processing may also be enabled and disabled selectively and dynamically. For example, the full-CSI processing may be enabled for a particular data transmission and disabled for some other data transmissions. The full-CSI processing may be enabled under certain conditions, for example, when the communication link has adequate SNR.

## MIMO Transmitter System with OFDM

[1045] FIG. 3 is a block diagram of an embodiment of a MIMO transmitter system 110c, which utilizes OFDM and is capable of adjusting its processing based on full or partial CSI. The information bits are encoded, interleaved, punctured, and symbol mapped by a TX data processor 114 to generate modulation symbols. The coding and modulation may be adjusted based on the available full or partial CSI reported by the receiver system. For a MIMO system with OFDM, the modulation symbols may be transmitted on multiple frequency subchannels and from multiple transmit antennas. When operating in a pure MIMO communication mode, the transmission on each frequency subchannel and from each transmit antenna represents non-duplicated data.

[1046] Within a MIMO processor 120c, a demultiplexer (DEMUX) 310 receives and demultiplexes the modulation symbols into a number of subchannel symbol streams, $S_1$ through $S_L$, one subchannel symbol stream for each frequency subchannel used to transmit the symbols.

[1047] For full-CSI processing, each subchannel symbol stream is then provided to a respective subchannel MIMO processor 312. Each subchannel MIMO processor 312 demultiplexes the received subchannel symbol stream into a number of (up to $N_C$) symbol substreams, one symbol substream for each spatial subchannel used to transmit the modulation symbols. For full-CSI processing in an OFDM system, the eigenmodes are derived and applied on a per frequency subchannel basis. Thus, each subchannel MIMO processors 312 preconditions up to $N_C$ modulation symbols in accordance with equation (1) to generate preconditioned modulation symbols. Each preconditioned modulation symbol for a particular transmit antenna of a particular frequency subchannel represents a linear combination of (weighted) modulation symbols for up to $N_C$ spatial subchannels.

[1048] For full-CSI processing, the (up to) $N_T$ preconditioned modulation symbols generated by each subchannel MIMO processor 312 for each time slot are

demultiplexed by a respective demultiplexer 314 and provided to (up to) $N_T$ symbol combiners 316a through 316t. For example, subchannel MIMO processor 312a assigned to frequency subchannel 1 may provide up to $N_T$ preconditioned modulation symbols for frequency subchannel 1 of antennas 1 through $N_T$. Similarly, subchannel MIMO processor 312*l* assigned to frequency subchannel L may provide up to $N_T$ symbols for frequency subchannel L of antennas 1 through $N_T$.

**[1049]** And for partial-CSI processing, each subchannel symbol stream, S, is demultiplexed by a respective demultiplexer 314 and provided to (up to) $N_T$ symbol combiners 316a through 316t. The processing by subchannel MIMO processor 312 is bypassed for partial-CSI processing.

**[1050]** Each combiner 316 receives the modulation symbols for up to L frequency subchannels, combines the symbols for each time slot into a modulation symbol vector V, and provides the modulation symbol vector to the next processing stage (i.e., modulator 122).

**[1051]** MIMO processor 120c thus receives and processes the modulation symbols to provide $N_T$ modulation symbol vectors, $V_1$ through $V_T$, one modulation symbol vector for each transmit antenna. Each modulation symbol vector V covers a single time slot, and each element of the modulation symbol vector V is associated with a specific frequency subchannel having a unique subcarrier on which the modulation symbol is conveyed. If not operating in a "pure" MIMO communication mode, some of the modulation symbol vectors may have duplicate or redundant information on specific frequency subchannels for different transmit antennas.

**[1052]** FIG. 3 also shows an embodiment of modulator 122 for OFDM. The modulation symbol vectors $V_1$ through $V_T$ from MIMO processor 120c are provided to modulators 122a through 122t, respectively. In the embodiment shown in FIG. 3, each modulator 122 includes an inverse Fast Fourier Transform (IFFT) 320, cycle prefix generator 322, and an upconverter 324.

**[1053]** IFFT 320 converts each received modulation symbol vector into its time-domain representation (which is referred to as an OFDM symbol) using IFFT. IFFT 320 can be designed to perform the IFFT on any number of frequency subchannels (e.g., 8, 16, 32, and so on). In an embodiment, for each modulation symbol vector converted to an OFDM symbol, cycle prefix generator 322 repeats a portion of the time-domain representation of the OFDM symbol to form a transmission symbol for a specific transmit antenna. The cyclic prefix insures that the transmission symbol retains its

orthogonal properties in the presence of multipath delay spread, thereby improving performance against deleterious path effects. The implementation of IFFT 320 and cycle prefix generator 322 is known in the art and not described in detail herein.

**[1054]** The time-domain representations from each cycle prefix generator 322 (i.e., the transmission symbols for each antenna) are then processed (e.g., converted into an analog signal, modulated, amplified, and filtered) by upconverter 324 to generate a modulated signal, which is then transmitted from the respective antenna 124.

**[1055]** OFDM modulation is described in further detail in a paper entitled "Multicarrier Modulation for Data Transmission : An Idea Whose Time Has Come," by John A.C. Bingham, IEEE Communications Magazine, May 1990, which is incorporated herein by reference.

**[1056]** A number of different types of transmission (e.g., voice, signaling, data, pilot, and so on) may be transmitted by a communication system. Each of these transmissions may require different processing.

**[1057]** FIG. 4 is a block diagram of a portion of a MIMO transmitter system 110d capable of providing different processing for different transmission types and which also employs OFDM. The aggregate input data, which includes all information bits to be transmitted by system 110d, is provided to a demultiplexer 408. Demultiplexer 408 demultiplexes the input data into a number of (K) channel data streams, $B_1$ through $B_K$. Each channel data stream may correspond to, for example, a signaling channel, a broadcast channel, a voice call, or a packet data transmission. Each channel data stream is provided to a respective TX data processor 114 that encodes the data using a particular encoding scheme selected for that channel data stream, interleaves the encoded data based on a particular interleaving scheme, and maps the interleaved bits into modulation symbols for one or more transmission channels used for transmitting that channel data stream.

**[1058]** The encoding can be performed on a per transmission basis (i.e., on each channel data stream, as shown in FIG. 4). However, the encoding may also be performed on the aggregate input data (as shown in FIG. 1), on a number of channel data streams, on a portion of a channel data stream, across a set of frequency subchannels, across a set of spatial subchannels, across a set of frequency subchannels and spatial subchannels, across each frequency subchannel, on each modulation symbol, or on some other unit of time, space, and frequency.

[1059] The modulation symbol stream from each TX data processor 114 may be transmitted on one or more frequency subchannels and via one or more spatial subchannels of each frequency subchannel. A TX MIMO processor 120d receives the modulation symbol streams from TX data processors 114. Depending on the communication mode to be used for each modulation symbol stream, TX MIMO processor 120d may demultiplex the modulation symbol stream into a number of subchannel symbol streams. In the embodiment shown in FIG. 4, modulation symbol stream $S_1$ is transmitted on one frequency subchannel and modulation symbol stream $S_K$ is transmitted on L frequency subchannels. The modulation stream for each frequency subchannel is processed by a respective subchannel MIMO processor 412, demultiplexed by demultiplexer 414, and combined by combiner 416 (e.g., in similar manner as that described in FIG. 3) to form a modulation symbol vector for each transmit antenna.

[1060] In general, a transmitter system codes and modulates data for each transmission channel based on information descriptive of that channel's transmission capability. This information is typically in the form of full CSI or partial CSI described above. The full/partial-CSI for the transmission channels used for data transmission is typically determined at the receiver system and reported back to the transmitter system, which then uses the information to adjust the coding and modulation accordingly. The techniques described herein are applicable for multiple parallel transmission channels supported by MIMO, OFDM, or any other communication scheme (e.g., a CDMA scheme) capable of supporting multiple parallel transmission channels.

[1061] MIMO processing is described in further detail in U.S Patent Application Serial No. 09/532,492, entitled "HIGH EFFICIENCY, HIGH PERFORMANCE COMMUNICATIONS SYSTEM EMPLOYING MULTI-CARRIER MODULATION," filed March 22, 2000, assigned to the assignee of the present application and incorporated herein by reference.

## MIMO Receiver System

[1062] Aspects of the invention provide techniques to process the received signals in a MIMO system to recover the transmitted data, and to estimate the characteristics of the MIMO channel. The estimated channel characteristics may then be reported back to the transmitter system and used to adjust the signal processing (e.g., coding, modulation, and so on). In this manner, high performance is achieved based on the

determined channel conditions. The receiver processing techniques described herein include a channel correlation matrix inversion (CCMI) technique, an unbiased minimum mean square error (UMMSE) technique, and a full-CSI technique, all of which are described in further detail below. Other receiver processing techniques may also be used and are within the scope of the invention.

**[1063]** FIG. 1 shows receiver system 150 having multiple ($N_R$) receive antennas and capable of processing a data transmission. The transmitted signals from up to $N_T$ transmit antennas are received by each of $N_R$ antennas 152a through 152r and routed to a respective demodulator (DEMOD) 154 (which is also referred to as a front-end processor). For example, receive antenna 152a may receive a number of transmitted signals from a number of transmit antennas, and receive antenna 152r may similarly receive multiple transmitted signals. Each demodulator 154 conditions (e.g., filters and amplifies) the received signal, downconverts the conditioned signal to an intermediate frequency or baseband, and digitizes the downconverted signal. Each demodulator 154 may further demodulate the digitized samples with a received pilot to generate received modulation symbols, which are provided to RX MIMO processor 156.

**[1064]** If OFDM is employed for the data transmission, each demodulator 154 further performs processing complementary to that performed by modulator 122 shown in FIG. 3. In this case, each demodulator 154 includes an FFT processor (not shown) that generates transformed representations of the samples and provides a stream of modulation symbol vectors, with each vector including L modulation symbols for L frequency subchannels. The modulation symbol vector streams from the FFT processors of all demodulators are then provided to a demultiplexer/combiner (not shown in FIG. 5), which first "channelizes" the modulation symbol vector stream from each FFT processor into a number of (up to L) subchannel symbol streams. Each of (up to) L subchannel symbol streams may then be provided to a respective RX MIMO processor 156.

**[1065]** For a MIMO system not utilizing OFDM, one RX MIMO processor 156 may be used to perform the MIMO processing for the modulation symbols from the $N_R$ received antennas. And for a MIMO system utilizing OFDM, one RX MIMO processor 156 may be used to perform the MIMO processing for the modulation symbols from the $N_R$ received antennas for each of the L frequency subchannels used for data transmission.

**[1066]** In a MIMO system with $N_T$ transmit antennas and $N_R$ receive antennas, the received signals at the output of the $N_R$ receive antennas may be expressed as:

$$\underline{\mathbf{r}} = \mathbf{H}\underline{\mathbf{x}} + \underline{\mathbf{n}} \quad , \qquad \text{Eq (2)}$$

where $\underline{\mathbf{r}}$ is the received symbol vector (i.e., the $N_R$ x 1 vector output from the MIMO channel, as measured at the receive antennas), $\mathbf{H}$ is the $N_R$ x $N_T$ channel coefficient matrix that gives the channel response for the $N_T$ transmit antennas and $N_R$ receive antennas at a specific time, $\underline{\mathbf{x}}$ is the transmitted symbol vector (i.e., the $N_T$ x 1 vector input into the MIMO channel), and $\underline{\mathbf{n}}$ is an $N_R$ x 1 vector representing noise plus interference. The received symbol vector $\underline{\mathbf{r}}$ includes $N_R$ modulation symbols from $N_R$ signals received via $N_R$ receive antennas at a specific time. Similarly, the transmitted symbol vector $\underline{\mathbf{x}}$ includes $N_T$ modulation symbols in $N_T$ signals transmitted via $N_T$ transmit antennas at a specific time.

## MIMO Receiver Utilizing CCMI Technique

**[1067]** For the CCMI technique, the receiver system first performs a channel matched filter operation on the received symbol vector $\underline{\mathbf{r}}$, and the filtered output can be expressed as:

$$\mathbf{H}^H \underline{\mathbf{r}} = \mathbf{H}^H \mathbf{H}\underline{\mathbf{x}} + \mathbf{H}^H \underline{\mathbf{n}} \quad , \qquad \text{Eq (3)}$$

where the superscript "$^H$" represents transpose and complex conjugate. A square matrix $\mathbf{R}$ may be used to denote the product of the channel coefficient matrix $\mathbf{H}$ with its conjugate-transpose $\mathbf{H}^H$ (i.e., $\mathbf{R} = \mathbf{H}^H \mathbf{H}$).

**[1068]** The channel coefficient matrix $\mathbf{H}$ may be derived, for example, from pilot symbols transmitted along with the data. In order to perform optimal reception and to estimate the SNR of the transmission channels, it is often convenient to insert some known symbols into the transmit data stream and to transmit the known symbols over one or more transmission channels. Such known symbols are also referred to as pilot symbols or pilot signals. Methods for estimating a single transmission channel based on a pilot signal or the data transmission may be found in a number of papers available in the art. One such channel estimation method is described by F. Ling in a paper entitled "Optimal Reception, Performance Bound, and Cutoff-Rate Analysis of References-Assisted Coherent CDMA Communications with Applications," IEEE Transaction On

Communication, Oct. 1999. This or some other channel estimation method may be extended to matrix form to derive the channel coefficient matrix **H**.

[1069] An estimate of the transmitted symbol vector, $\underline{\mathbf{x}}'$, may be obtained by multiplying the signal vector $\mathbf{H}^H\underline{\mathbf{r}}$ with the inverse (or pseudo-inverse) of **R**, which can be expressed as:

$$\begin{aligned}\underline{\mathbf{x}}' &= \mathbf{R}^{-1}\mathbf{H}^H\underline{\mathbf{r}} \\ &= \underline{\mathbf{x}} + \mathbf{R}^{-1}\mathbf{H}^H\underline{\mathbf{n}} \\ &= \underline{\mathbf{x}} + \underline{\mathbf{n}}' \quad .\end{aligned} \qquad \text{Eq (4)}$$

From the above equations, it can be observed that the transmitted symbol vector $\underline{\mathbf{x}}$ may be recovered by matched filtering (i.e., multiplying with the matrix $\mathbf{H}^H$) the received symbol vector $\underline{\mathbf{r}}$ and then multiplying the filtered result with the inverse square matrix $\mathbf{R}^{-1}$.

[1070] The SNR of the transmission channels may be determined as follows. The autocorrelation matrix $\phi_{nn}$ of the noise vector $\underline{\mathbf{n}}$ is first computed from the received signal. In general, $\phi_{nn}$ is a Hermitian matrix, i.e., it is complex-conjugate-symmetric. If the components of the channel noise are uncorrelated and further independent and identically distributed (iid), the autocorrelation matrix $\phi_{nn}$ of the noise vector $\underline{\mathbf{n}}$ can be expressed as:

$$\phi_{nn} = \sigma_n^2\mathbf{I} \text{ , and} \qquad \text{Eq (5)}$$

$$\phi_{nn}^{-1} = \frac{1}{\sigma_n^2}\mathbf{I} \text{ ,}$$

where **I** is the identity matrix (i.e., ones along the diagonal and zeros otherwise) and $\sigma_n^2$ is the noise variance of the received signals. The autocorrelation matrix $\phi_{n'n'}$ of the post-processed noise vector $\underline{\mathbf{n}}'$ (i.e., after the matched filtering and pre-multiplication with the matrix $\mathbf{R}^{-1}$) can be expressed as:

$$\begin{aligned}\phi_{\mathbf{n}'\mathbf{n}'} &= E[\underline{\mathbf{n}}'\underline{\mathbf{n}}'^H] \\ &= \sigma_n^2\mathbf{R}^{-1}\end{aligned} \qquad \text{Eq (6)}$$

From equation (6), the noise variance $\sigma_{n'}^2$ of the i-th element of the post-processed noise $\underline{\mathbf{n}}'$ is equal to $\sigma_n^2\tilde{r}_{ii}$, where $\tilde{r}_{ii}$ is the i-th diagonal element of $\mathbf{R}^{-1}$. For a MIMO system

not utilizing OFDM, the i-th element is representative of the i-th receive antenna. And if OFDM is utilized, then the subscript "i" may be decomposed into a subscript "jk", where "j" represents the j-th frequency subchannel and "k" represents the k-th spatial subchannel corresponding to the k-th receive antenna.

**[1071]** For the CCMI technique, the SNR of the i-th element of the received symbol vector after processing (i.e., the i-th element of $\underline{\mathbf{x}}'$) can be expressed as:

$$SNR_i = \frac{\overline{|x'_i|^2}}{\sigma_{n'}^2} \ . \qquad \text{Eq (7)}$$

If the variance of the i-th transmitted symbol $\overline{|x'_i|^2}$ is equal to one (1.0) on the average, the SNR of the receive symbol vector may be expressed as:

$$SNR_i = \frac{1}{\tilde{r}_{ii}\sigma_n^2} \ .$$

The noise variance may be normalized by scaling the i-th element of the received symbol vector by $1/\sqrt{\tilde{r}_{ii}}$.

**[1072]** The scaled signals from the $N_R$ receive antennas may be summed together to form a combined signal, which may be expressed as:

$$x'_{total} = \sum_{i=1}^{N_R} \frac{x'_i}{\tilde{r}_{ii}} \ . \qquad \text{Eq (8)}$$

The SNR of the combined signal, $SNR_{total}$, would then have a maximal combined SNR that is equal to the sum of the SNR of the signals from the $N_R$ receive antennas. The combined SNR may be expressed as:

$$SNR_{total} = \sum_{i=1}^{N_R} SNR_i = \frac{1}{\sigma_n^2} \sum_{i=1}^{N_R} \frac{1}{\tilde{r}_{ii}} \ . \qquad \text{Eq (9)}$$

**[1073]** FIG. 5 shows an embodiment of an RX MIMO processor 156a, which is capable of implementing the CCMI processing described above. Within RX MIMO processor 156a, the modulation symbols from the $N_R$ receive antennas are multiplexed by a multiplexer 512 to form a stream of received modulation symbol vectors $\underline{\mathbf{r}}$. The channel coefficient matrix $\mathbf{H}$ may be estimated based on pilot signals similar to conventional pilot assisted single and multi-carrier systems, as is known in the art. The matrix $\mathbf{R}$ is then computed according to $\mathbf{R} = \mathbf{H}^H\mathbf{H}$ as shown above. The received modulation symbol vectors $\underline{\mathbf{r}}$ are then filtered by a match filter 514, which pre-

multiplies each vector $\underline{\mathbf{r}}$ with the conjugate-transpose channel coefficient matrix $\mathbf{H}^H$, as shown above in equation (3). The filtered vectors are further pre-multiplied by a multiplier 516 with the inverse square matrix $\mathbf{R}^{-1}$ to form an estimate $\underline{\mathbf{x}}'$ of the transmitted modulation symbol vector $\underline{\mathbf{x}}$, as shown above in equation (4).

**[1074]** For certain communication modes, the subchannel symbol streams from all antennas used for the transmission of the channel data stream may be provided to a combiner 518, which combines redundant information across time, space, and frequency. The combined modulation symbols $\underline{\mathbf{x}}''$ are then provided to RX data processor 158. For some other communication modes, the estimated modulation symbols $\underline{\mathbf{x}}'$ may be provided directly to RX data processor 158 (not shown in FIG. 5).

**[1075]** RX MIMO processor 156a thus generates a number of independent symbol streams corresponding to the number of transmission channels used at the transmitter system. Each symbol stream includes post-processed modulation symbols, which correspond to the modulation symbols prior to the full/partial-CSI processing at the transmitter system. The (post-processed) symbol streams are then provided to RX data processor 158.

**[1076]** Within RX data processor 158, each post-processed symbol stream of modulation symbols is provided to a respective demodulation element that implements a demodulation scheme (e.g., M-PSK, M-QAM) that is complementary to the modulation scheme used at the transmitter system for the transmission channel being processed. For the MIMO communication mode, the demodulated data from all assigned demodulators may then be decoded independently or multiplexed into one channel data stream and then decoded, depending upon the coding and modulation method employed at the transmitter unit. Each channel data stream may then be provided to a respective decoder that implements a decoding scheme complementary to that used at the transmitter unit for the channel data stream. The decoded data from each decoder represents an estimate of the transmitted data for that channel data stream.

**[1077]** The estimated modulation symbols $\underline{\mathbf{x}}'$ and/or the combined modulation symbols $\underline{\mathbf{x}}''$ are also provided to a CSI processor 520, which determines full or partial CSI for the transmission channels and provides the full/partial-CSI to be reported back to transmitter system 110. For example, CSI processor 520 may estimate the noise covariance matrix $\phi_{nn}$ of the i-th transmission channel based on the received pilot signal and then compute the SNR based on equations (7) and (9). The SNR can be estimated

similar to conventional pilot assisted single and multi-carrier systems, as is known in the art. The SNR for the transmission channels comprises the partial-CSI that is reported back to the transmitter system. The modulation symbols are further provided to a channel estimator 522 and a matrix processor 524 that respectively estimates the channel coefficient matrix **H** and derive the square matrix **R**. A controller 530 couples to RX MIMO processor 156a and RX data processor 158 and directs the operation of these units.

## MIMO Receiver Utilizing UMMSE Technique

**[1078]** For the UMMSE technique, the receiver system performs a multiplication of the received symbol vector $\underline{\mathbf{r}}$ with a matrix **M** to derive an initial MMSE estimate $\hat{\underline{\mathbf{x}}}$ of the transmitted symbol vector $\underline{\mathbf{x}}$, which can be expressed as:

$$\hat{\underline{\mathbf{x}}} = \mathbf{M}\underline{\mathbf{r}} \ . \qquad \text{Eq (10)}$$

The matrix **M** is selected such that the mean square error of the error vector $\underline{\mathbf{e}}$ between the initial MMSE estimate $\hat{\underline{\mathbf{x}}}$ and the transmitted symbol vector $\underline{\mathbf{x}}$ (i.e., $\mathbf{e} = \hat{\mathbf{x}} - \mathbf{x}$) is minimized.

**[1079]** To determine **M**, a cost function $\varepsilon$ can first be expressed as:

$$\begin{aligned} \varepsilon &= E\{\underline{\mathbf{e}}^H \underline{\mathbf{e}}\} \\ &= E\{[\underline{\mathbf{r}}^H \mathbf{M}^H - \underline{\mathbf{x}}^H][\mathbf{M}\underline{\mathbf{r}} - \underline{\mathbf{x}}]\} \\ &= E\{\underline{\mathbf{r}}^H \mathbf{M}^H \mathbf{M}\underline{\mathbf{r}} - 2\mathrm{Re}[\underline{\mathbf{x}}^H \mathbf{M}\underline{\mathbf{r}}] + \underline{\mathbf{x}}^H \underline{\mathbf{x}}\} \quad . \end{aligned}$$

To minimize the cost function $\varepsilon$, a derivative of the cost function can be taken with respect to **M**, and the result can be set to zero, as follows:

$$\frac{\partial}{\partial \mathbf{M}} \varepsilon = 2(\mathbf{H}\mathbf{H}^H + \phi_{nn})\mathbf{M}^H - 2\mathbf{H} = \mathbf{0} \ .$$

Using the equalities $E\{\underline{\mathbf{x}}\underline{\mathbf{x}}^H\} = \mathbf{I}$, $E\{\underline{\mathbf{r}}\underline{\mathbf{r}}^H\} = \mathbf{H}\mathbf{H}^H + \phi_{nn}$, and $E\{\underline{\mathbf{r}}\underline{\mathbf{x}}^H\} = \mathbf{H}$, the following is obtained:

$$2(\mathbf{H}\mathbf{H}^H + \phi_{nn})\mathbf{M}^H = 2\mathbf{H} \ .$$

Thus, the matrix **M** can be expressed as:

$$\mathbf{M} = \mathbf{H}^H (\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1} \ . \qquad \text{Eq (11)}$$

[1080] Based on equations (10) and (11), the initial MMSE estimate $\hat{\underline{\mathbf{x}}}$ of the transmitted symbol vector $\underline{\mathbf{x}}$ can be determined as:

$$\begin{aligned}\hat{\underline{\mathbf{x}}} &= \mathbf{M}\underline{\mathbf{r}} \\ &= \mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1}\underline{\mathbf{r}} \quad .\end{aligned} \qquad \text{Eq (12)}$$

[1081] To determine the SNR of the transmission channels for the UMMSE technique, the signal component can first be determined based on the mean of $\hat{\underline{\mathbf{x}}}$ given $\underline{\mathbf{x}}$, averaged over the additive noise, which can be expressed as:

$$\begin{aligned}E[\hat{\underline{\mathbf{x}}} \mid \underline{\mathbf{x}}] &= E[\mathbf{M}\underline{\mathbf{r}} \mid \underline{\mathbf{x}}] \\ &= \mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1}E[\underline{\mathbf{r}}] \\ &= \mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1}\mathbf{H}\underline{\mathbf{x}} \\ &= \mathbf{V}\underline{\mathbf{x}} \quad ,\end{aligned}$$

where the matrix $\mathbf{V}$ is defined as:

$$\begin{aligned}\mathbf{V} &= \{v_{ij}\} \\ &= \mathbf{M}\mathbf{H} \\ &= \mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1}\mathbf{H} \quad .\end{aligned}$$

Using the identity

$$(\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1} = \phi_{nn}^{-1} - \phi_{nn}^{-1}\mathbf{H}(\mathbf{I} + \mathbf{H}^H\phi_{nn}^{-1}\mathbf{H})^{-1}\mathbf{H}^H\phi_{nn}^{-1} \quad ,$$

the matrix $\mathbf{V}$ can be expressed as:

$$\mathbf{V} = \mathbf{H}^H\phi_{nn}^{-1}\mathbf{H}(\mathbf{I} + \mathbf{H}^H\phi_{nn}^{-1}\mathbf{H})^{-1} \quad .$$

[1082] The i-th element of the initial MMSE estimate $\hat{\underline{\mathbf{x}}}$, $\hat{x}_i$, can be expressed as:

$$\hat{x}_i = v_{i1}x_1 + \ldots + v_{ii}x_i + \ldots + v_{iN_R}x_{N_R} \quad . \qquad \text{Eq (13)}$$

If all of the elements of $\hat{\underline{\mathbf{x}}}$ are uncorrelated and have zero mean, the expected value of the i-th element of $\hat{\underline{\mathbf{x}}}$ can be expressed as:

$$E[\hat{x}_i \mid \mathbf{x}] = v_{ii}x_i \quad . \qquad \text{Eq (14)}$$

**[1083]** As shown in equation (14), $\hat{x}_i$ is a biased estimate of $x_i$. This bias can be removed to obtain improved receiver performance in accordance with the UMMSE technique. An unbiased estimate of $x_i$ can be obtained by dividing $\hat{x}_i$ by $v_{ii}$. Thus, the unbiased minimum mean square error estimate of $\underline{\mathbf{x}}$, $\underline{\tilde{\mathbf{x}}}$, can be obtained by pre-multiplying the biased estimate $\underline{\hat{\mathbf{x}}}$ by a diagonal matrix $\mathbf{D}_{\mathbf{V}}^{-1}$, as follows:

$$\underline{\tilde{\mathbf{x}}} = \mathbf{D}_{\mathbf{V}}^{-1}\underline{\hat{\mathbf{x}}} \quad , \qquad \text{Eq (15)}$$

where

$$\mathbf{D}_{\mathbf{V}}^{-1} = diag(1/v_{11}, 1/v_{22}, \ldots, 1/v_{N_R N_R}) \ .$$

**[1084]** To determine the noise plus interference, the error $\underline{\hat{\mathbf{e}}}$ between the unbiased estimate $\underline{\tilde{\mathbf{x}}}$ and the transmitted symbol vector $\underline{\mathbf{x}}$ can be expressed as:

$$\begin{aligned}\underline{\hat{\mathbf{e}}} &= \underline{\mathbf{x}} - \mathbf{D}_{\mathbf{V}}^{-1}\underline{\hat{\mathbf{x}}} \\ &= \underline{\mathbf{x}} - \mathbf{D}_{\mathbf{V}}^{-1}\mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1}\underline{\mathbf{r}} \ .\end{aligned}$$

The autocorrelation matrix of the error vector $\underline{\hat{\mathbf{e}}}$ can be expressed as:

$$\begin{aligned}\phi_{\hat{e}\hat{e}} &\cong \mathbf{U} \cong \{u_{ij}\} = E[\hat{\mathbf{e}}\hat{\mathbf{e}}^H] \\ &= \mathbf{I} - \mathbf{D}_{\mathbf{V}}^{-1}\mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1}\mathbf{H}(1 - \frac{1}{2}\mathbf{D}_{\mathbf{V}}^{-1}) - (1 - \frac{1}{2}\mathbf{D}_{\mathbf{V}}^{-1})\mathbf{H}^H(\mathbf{H}\mathbf{H}^H + \phi_{nn})^{-1}\mathbf{H}\mathbf{D}_{\mathbf{V}}^{-1} .\end{aligned}$$

The variance of the i-th element of the error vector $\underline{\hat{\mathbf{e}}}$ is equal to $u_{ii}$. The elements of the error vector $\underline{\hat{\mathbf{e}}}$ are correlated. However, sufficient interleaving may be used such that the correlation between the elements of the error vector $\underline{\hat{\mathbf{e}}}$ can be ignored and only the variance affects system performance.

**[1085]** If the components of the channel noise are uncorrelated and iid, the correlation matrix of the channel noise can be expressed as shown in equation (5). In that case, the autocorrelation matrix of the error vector $\underline{\hat{\mathbf{e}}}$ can be expressed as:

$$\begin{aligned}\phi_{\hat{e}\hat{e}} &= \mathbf{I} - \mathbf{D}_{\mathbf{X}}^{-1}[\mathbf{I} - \sigma_n^2(\sigma_n^2\mathbf{I} + \mathbf{R})^{-1}](\mathbf{I} - \frac{1}{2}\mathbf{D}_{\mathbf{X}}^{-1}) - (\mathbf{I} - \frac{1}{2}\mathbf{D}_{\mathbf{X}}^{-1})[\mathbf{I} - \sigma_n^2(\sigma_n^2\mathbf{I} + \mathbf{R})^{-1}]\mathbf{D}_{\mathbf{X}}^{-1} \\ &= \mathbf{U} = \{u_{ij}\} .\end{aligned} \qquad \text{Eq (16)}$$

And if the components of the channel noise are uncorrelated, then

$$\mathbf{U} = \mathbf{I} - \mathbf{D}_{\mathbf{V}}^{-1}\mathbf{H}^{H}(\mathbf{H}\mathbf{H}^{H} + \phi_{nn})^{-1}\mathbf{H}(\mathbf{I} - \frac{1}{2}\mathbf{D}_{\mathbf{V}}^{-1}) - (\mathbf{I} - \frac{1}{2}\mathbf{D}_{\mathbf{V}}^{-1})\mathbf{H}^{H}(\mathbf{H}\mathbf{H}^{H} + \phi_{nn})^{-1}\mathbf{H}\mathbf{D}_{\mathbf{V}}^{-1}. \qquad \text{Eq (17)}$$

The SNR of the demodulator output corresponding to the i-th transmitted symbol can be expressed as:

$$SNR_i = \frac{E[\overline{|x_i|^2}]}{u_{ii}} \quad . \qquad \text{Eq (18)}$$

If the variance, $\overline{|x_i|^2}$, of the processed received symbols, $x_i$, is equal to one (1.0) on the average, the SNR of for the receive symbol vector may be expressed as:

$$SNR_i = \frac{1}{u_{ii}} \quad .$$

**[1086]** FIG. 6 shows an embodiment of an RX MIMO processor 156b, which is capable of implementing the UMMSE processing described above. Similar to the CCMI method, the matrices **H** and $\phi_{nn}$ may be first estimated based on the received pilot signals and/or data transmissions. The weighting coefficient matrix **M** is then computed according to equation (11). Within RX MIMO processor 156b, the modulation symbols from the $N_R$ receive antennas are multiplexed by a multiplexer 612 to form a stream of received modulation symbol vectors $\underline{\mathbf{r}}$. The received modulation symbol vectors $\underline{\mathbf{r}}$ are then pre-multiplied by a multiplier 614 with the matrix **M** to form an estimate $\hat{\underline{\mathbf{x}}}$ of the transmitted symbol vector $\underline{\mathbf{x}}$, as shown above in equation (10). The estimate $\hat{\underline{\mathbf{x}}}$ is further pre-multiplied by a multiplier 616 with the diagonal matrix $\mathbf{D}_{\mathbf{V}}^{-1}$ to form an unbiased estimate $\tilde{\underline{\mathbf{x}}}$ of the transmitted symbol vector $\underline{\mathbf{x}}$, as shown above in equation (15).

**[1087]** Again, depending on the particular communication mode being implemented, the subchannel symbol streams from all antennas used for the transmission of the channel data stream may be provided to a combiner 618, which combines redundant information across time, space, and frequency. The combined modulation symbols $\tilde{\underline{\mathbf{x}}}''$ are then provided to RX data processor 158. And for some other communication modes, the estimated modulation symbols $\tilde{\underline{\mathbf{x}}}$ may be provided directly to RX data processor 158.

**[1088]** The unbiased estimated modulation symbols $\underline{\tilde{\mathbf{x}}}$ and/or the combined modulation symbols $\underline{\tilde{\mathbf{x}}}''$ are also provided to a CSI processor 620, which determines full or partial CSI for the transmission channels and provides the full/partial-CSI to be reported back to transmitter system 110. For example, CSI processor 620 may estimate the SNR of the i-th transmission channel according to equations (16) through (18). The SNR for the transmission channels comprises the partial-CSI that is reported back to the transmitter system. The optimal M as computed in equation (11) should already minimize the norm of the error vector. $\mathbf{D_v}$ is computed in accordance with equation (16).

## MIMO Receiver Utilizing Full-CSI Technique

**[1089]** For the full-CSI technique, the received signals at the output of the $N_R$ receive antennas may be expressed as shown above in equation (2), which is:

$$\underline{\mathbf{r}} = \mathbf{H}\underline{\mathbf{x}} + \underline{\mathbf{n}} \quad .$$

The eigenvector decomposition of the Hermitian matrix formed by the product of the channel matrix with its conjugate-transpose can be expressed as:

$$\mathbf{H}^H\mathbf{H} = \mathbf{E}\boldsymbol{\Lambda}\mathbf{E}^H \quad ,$$

where E is the eigenvector matrix, and $\boldsymbol{\Lambda}$ is a diagonal matrix of eigenvalues, both of dimension $N_T x N_T$. The transmitter preconditions a set of $N_T$ modulation symbols $\underline{\mathbf{b}}$ using the eigenvector matrix $\mathbf{E}$, as shown above in equation (1). The transmitted (preconditioned) modulation symbols from the $N_T$ transmit antennas can thus be expressed as:

$$\underline{\mathbf{x}} = \mathbf{E}\underline{\mathbf{b}} \quad .$$

Since $\mathbf{H}^H\mathbf{H}$ is Hermitian, the eigenvector matrix is unitary. Thus, if the elements of $\underline{\mathbf{b}}$ have equal power, the elements of $\underline{\mathbf{x}}$ also have equal power. The received signal may then be expressed as:

$$\underline{\mathbf{r}} = \mathbf{H}\mathbf{E}\underline{\mathbf{b}} + \underline{\mathbf{n}} \quad . \qquad \text{Eq (19)}$$

**[1090]** The receiver performs a channel-matched-filter operation, followed by multiplication by the right eigenvectors. The result of the channel-matched-filter and multiplication operations is a vector $\underline{\mathbf{z}}$, which can be expressed as:

$$\underline{\mathbf{z}} = \mathbf{E}^H\mathbf{H}^H\mathbf{H}\mathbf{E}\underline{\mathbf{b}} + \mathbf{E}^H\mathbf{H}^H\underline{\mathbf{n}} = \boldsymbol{\Lambda}\underline{\mathbf{b}} + \underline{\mathbf{n}}' \quad , \qquad \text{Eq (20)}$$

where the new noise term has covariance that can be expressed as:

$$E(\underline{\hat{n}}\underline{\hat{n}}^H) = E(\mathbf{E}^H\mathbf{H}^H\underline{\mathbf{n}}\underline{\mathbf{n}}^H\mathbf{H}\mathbf{E}) = \mathbf{E}^H\mathbf{H}^H\mathbf{H}\mathbf{E} = \Lambda \quad , \qquad \text{Eq (21)}$$

i.e., the noise components are independent with variance given by the eigenvalues. The SNR of the i-th component of $\underline{\mathbf{z}}$ is $\lambda_i$, the i-th diagonal element of $\mathbf{\Lambda}$.

**[1091]** Full-CSI processing is described in further detail in the aforementioned U.S Patent Application Serial No. 09/532,492.

**[1092]** The receiver embodiment shown in FIG. 5 may also be used to implement the full-CSI technique. The received modulation symbol vectors $\underline{\mathbf{r}}$ are filtered by match filter 514, which pre-multiplies each vector $\underline{\mathbf{r}}$ with the conjugate-transpose channel coefficient matrix $\mathbf{H}^H$, as shown above in equation (20). The filtered vectors are further pre-multiplied by multiplier 516 with the right eigenvectors $\mathbf{E}^H$ to form an estimate $\underline{\mathbf{z}}$ of the modulation symbol vector $\underline{\mathbf{b}}$, as shown above in equation (20). For the full-CSI technique, matrix processor 524 is configured to provide the right eigenvectors $\mathbf{E}^H$. The subsequent processing (e.g., by combiner 518 and RX data processor 158) may be achieved as described above.

**[1093]** For the full-CSI technique, the transmitter unit can select a coding scheme and a modulation scheme (i.e., a signal constellation) for each of the eigenvectors based on the SNR that is given by the eigenvalue. Providing that the channel conditions do not change appreciably in the interval between the time the CSI is measured at the receiver and reported and used to precondition the transmission at the transmitter, the performance of the communications system may be equivalent to that of a set of independent AWGN channels with known SNRs.

## Reporting Full or Partial CSI back to the Transmitter System

**[1094]** Using either the partial-CSI (e.g., CCMI or UMMSE) or full-CSI technique described herein, the SNR of each transmission channel may be obtained for the received signals. The determined SNR for the transmission channels may then be reported back to the transmitter system via a reverse channel. By feeding back the SNR values of the transmitted modulation symbols for the transmission channels (i.e., for each spatial subchannel, and possibly for each frequency subchannel if OFDM is employed), it is possible to implement adaptive processing (e.g., adaptive coding and modulation) to improve utilization of the MIMO channel. For the partial-CSI feedback techniques, adaptive processing may be achieved without complete CSI. For the full-

CSI feedback techniques, sufficient information (and not necessarily the explicit eigenvalues and eignemodes) is fed back to the transmitter to facilitate calculation of the eigenvalues and eigenmodes for each frequency subchannel utilized.

[1095] For the CCMI technique, the SNR values of the received modulation symbols (e.g., $SNR_i = \overline{|x'_i|^2} / \sigma_{n'}^2$ or $SNR_i = 1/\sigma_n^2 \hat{r}_{ii}$ for the symbol received on the i-th transmission channel) are fed back to the transmitter. For the UMMSE technique, the SNR values of the received modulation symbols (e.g., $SNR_i = E[|x_i|^2]/u_{ii}$ or $SNR_i = 1/u_{ii}$ for the symbol received on the i-th transmission channel, with $u_{ii}$ being computed as shown above in equations (16) and (17)) are fed back to the transmitter. And for the full-CSI technique, the SNR values of the received modulation symbols (e.g., $SNR_i = \overline{|z_i|^2} / \sigma_{n'}^2$ or $SNR_i = \lambda_{ii} / \sigma_n^2$ for the symbol received on the i-th transmission channel, where $\lambda_{ii}$ is the eigenvalue of the square matrix **R**) can be fed back to the transmitter. For the full-CSI technique, the eigenmodes **E** may be further determined and fed back to the transmitter. For the partial and full-CSI techniques, the SNR are used at the transmitter system to adjust the processing of the data. And for the full-CSI technique, the eigenmodes **E** are further used to precondition the modulation symbols prior to transmission.

[1096] The CSI to be reported back to the transmitter may be sent in full, differentially, or a combination thereof. In one embodiment, full or partial CSI is reported periodically, and differential updates are sent based on the prior transmitted CSI. As an example for full CSI, the updates may be corrections (based on an error signal) to the reported eigenmodes. The eigenvalues typically do not change as rapidly as the eigenmodes, so these may be updated at a lower rate. In another embodiment, the CSI is sent only when there is a change (e.g., if the change exceeds a particular threshold), which may lower the effective rate of the feedback channel. As an example for partial CSI, the SNRs may be sent back (e.g., differentially) only when they change. For an OFDM system (with or without MIMO), correlation in the frequency domain may be exploited to permit reduction in the amount of CSI to be fed back. As an example for an OFDM system using partial CSI, if the SNR corresponding to a particular spatial subchannel for M frequency subchannels is the same, the SNR and the first and last frequency subchannels for which this condition is true may be reported. Other compression and feedback channel error recovery techniques to reduce the

amount of data to be fed back for CSI may also be used and are within the scope of the invention.

**[1097]** Referring back to FIG. 1, the full or partial-CSI (e.g., channel SNR) determined by RX MIMO processor 156 is provided to a TX data processor 162, which processes the CSI and provides processed data to one or more modulators 154. Modulators 154 further condition the processed data and transmit the CSI back to transmitter system 110 via a reverse channel.

**[1098]** At system 110, the transmitted feedback signal is received by antennas 124, demodulated by demodulators 122, and provided to a RX data processor 132. RX data processor 132 performs processing complementary to that performed by TX data processor 162 and recovers the reported full/partial-CSI, which is then provided to, and used to adjust the processing by, TX data processor 114 and TX MIMO processor 120.

**[1099]** Transmitter system 110 may adjust (i.e., adapt) its processing based on the full/partial-CSI (e.g., SNR information) from receiver system 150. For example, the coding for each transmission channel may be adjusted such that the information bit rate matches the transmission capability supported by the channel SNR. Additionally, the modulation scheme for the transmission channel may be selected based on the channel SNR. Other processing (e.g., interleaving) may also be adjusted and are within the scope of the invention. The adjustment of the processing for each transmission channel based on the determined SNR for the channel allows the MIMO system to achieve high performance (i.e., high throughput or bit rate for a particular level of performance). The adaptive processing can be applied to a single-carrier MIMO system or a multi-carrier based MIMO system (e.g., a MIMO system utilizing OFDM).

**[1100]** The adjustment in the coding and the selection of the modulation scheme at the transmitter system may be achieved based on numerous techniques, one of which is described in the aforementioned U.S Patent Application Serial No. 09/776,073.

**[1101]** The partial (e.g., CCMI and UMMSE) and full-CSI techniques are receiver processing techniques that allow a MIMO system to utilize the additional dimensionalities created by the use of multiple transmit and receive antennas, which is a main advantage for employing MIMO. The CCMI and UMMSE techniques may allow the same number of modulation symbols to be transmitted for each time slot as for a MIMO system utilizing full CSI. However, other receiver processing techniques may also be used in conjunction with the full/partial-CSI feedback techniques described herein and are within the scope of the invention. Analogously, FIGS. 5 and 6 represent

two embodiments of a receiver system capable of processing a MIMO transmission, determining the characteristics of the transmission channels (i.e., the SNR), and reporting full or partial CSI back to the transmitter system. Other designs based on the techniques presented herein and other receiver processing techniques can be contemplated and are within the scope of the invention.

**[1102]** The partial-CSI technique (e.g., CCMI and UMMSE techniques) may also be used in a straightforward manner without adaptive processing at the transmitter when only the overall received signal SNR or the attainable overall throughput estimated based on such SNR is feed back. In one implementation, a modulation format is determined based on the received SNR estimate or the estimated throughput, and the same modulation format is used for all transmission channels. This method may reduce the overall system throughput but may also greatly reduce the amount of information sent back over the reverse link.

**[1103]** Improvement in system performance may be realized with the use of the full/partial-CSI feedback techniques of the invention. The system throughput with partial CSI feedback can be computed and compared against the throughput with full CSI feedback. The system throughput can be defined as:

$$C = \sum_{i=1}^{N_C} \log_2(1+\gamma_i) \quad ,$$

where $\gamma_i$ is the SNR of each received modulation symbol for partial CSI techniques or the SNR of each transmission channel for the full CSI technique. The SNR for various processing techniques can be summarized as follows:

$$\gamma_i = \frac{1}{\sigma_n^2 r_{ii}} \;, \qquad \text{for the CCMI technique}$$

$$\gamma_i = \frac{1}{u_{ii}} \;, \qquad \text{for the UMMSE technique, and}$$

$$\gamma_i = \frac{\lambda_{ii}}{\sigma_n^2} \;, \qquad \text{for full CSI technique.}$$

**[1104]** FIGS. 7A and 7B show the performance of a 4 x 4 MIMO system employing partial-CSI and full-CSI feedback techniques. The results are obtained from a computer simulation. In the simulation, the elements of each channel coefficient matrix **H** are modeled as independent Gaussian random variable with zero mean and unity variance.

For each calculation, a number of random matrix realizations are generated and the throughput computed for the realization are averaged to generate the average throughput.

**[1105]** FIG. 7A shows the average throughput for the MIMO system for the full-CSI, partial-CSI CCMI, and partial-CSI UMMSE techniques for different SNR values. It can be seen from FIG. 7A that the throughput of the partial-CSI UMMSE technique is approximately 75% of the full- SI throughput at high SNR values, and approaches the full CSI throughput at low SNR values. The throughput of the partial-CSI CCMI technique is approximately 75%-90% of the throughput of the partial-CSI UMMSE technique at high SNR values, and is approximately less than 30% of the UMMSE throughput at low SNR values.

**[1106]** FIG. 7B shows the cumulative probability distribution functions (CDF) for the three techniques generated based on the histogram of the data. FIG. 7B shows that at an average SNR of 16 dB per transmission channel, there are approximately 5% cases when the throughput is less than 2 bps/Hz for the CCMI technique. On the other hand, the throughput of the UMMSE technique is above 7.5 bps/Hz for all cases at the same SNR. Thus, the UMMSE technique is likely to have lower outage probability than the CCMI technique.

**[1107]** The elements of the transmitter and receiver systems may be implemented with one or more digital signal processors (DSP), application specific integrated circuits (ASIC), processors, microprocessors, controllers, microcontrollers, field programmable gate arrays (FPGA), programmable logic devices, other electronic units, or any combination thereof. Some of the functions and processing described herein may also be implemented with software executed on a processor.

**[1108]** Aspects of the invention may be implemented with a combination of software and hardware. For example, computations for the symbol estimates for the CCMI and UMMSE techniques and the derivation of the channel SNR may be performed based on program codes executed on a processor (controllers 530 and 650 in FIGS. 5 and 6, respectively).

**[1109]** The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not

intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[1110] **WHAT IS CLAIMED IS:**

# CLAIMS

1. A method for transmitting data from a transmitter unit to a receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

at the receiver unit,

receiving a plurality of signals via a plurality of receive antennas, wherein the received signal from each receive antenna comprises a combination of one or more signals transmitted from the transmitter unit,

processing the received signals to derive channel state information (CSI) indicative of characteristics of a plurality of transmission channels used for data transmission, and

transmitting the CSI back to the transmitter unit; and

at the transmitter unit,

receiving the CSI from the receiver unit, and

processing data for transmission to the receiver unit based on the received CSI.

2. The method of claim 1, wherein the reported CSI comprises signal-to-noise-plus-interference (SNR) estimates for each of the plurality of transmission channels.

3. The method of claim 2, wherein the processing at the transmitter unit includes

coding data for each transmission channel based on the SNR estimate for the transmission channel.

4. The method of claim 3, wherein the data for each transmission channel is independently coded based on the SNR estimate for the transmission channel.

5. The method of claim 3, wherein the coding includes

coding the data for the transmission channel with a fixed base code, and

adjusting puncturing of coded bits based on the SNR estimate for the transmission channel.

6. The method of claim 3, wherein the processing at the transmitter unit further includes

modulating coded data for each transmission channel in accordance with a modulation scheme selected based on the SNR estimate for the transmission channel.

7. The method of claim 1, wherein the reported CSI comprises characterizations for the plurality of transmission channels.

8. The method of claim 1, wherein the reported CSI is indicative of eigenmodes and eigenvalues for the plurality of transmission channels.

9. The method of claim 8, wherein the processing at the transmitter unit includes

coding data for the transmission channels based on the eigenvalues.

10. The method of claim 9, wherein the data for each transmission channel is independently coded.

11. The method of claim 9, wherein the processing at the transmitter unit further includes

modulating coded data for the transmission channels in accordance with modulation schemes selected based on the eigenvalues to provide modulation symbols.

12. The method of claim 11, wherein the processing at the transmitter unit further includes

pre-conditioning the modulation symbols prior to transmission based on the eigenmodes.

13. The method of claim 1, wherein the CSI is transmitted in full from the receiver unit.

14. The method of claim 13, wherein the CSI is periodically transmitted in full from the receiver unit, and wherein updates to the CSI are transmitted between full transmissions.

15. The method of claim 1, wherein the CSI is transmitted when changes in the channel characteristics exceeding a particular threshold are detected.

16. The method of claim 8, wherein the CSI indicative of the eigenmodes and eigenvalues are transmitted at different update rates.

17. The method of claim 1, wherein the CSI is derived at the receiver unit based on a correlation matrix inversion (CCMI) processing.

18. The method of claim 17, wherein the CCMI processing at the receiver unit includes

processing the received signals to derive received modulation symbols;

filtering the received modulation symbols in accordance with a first matrix to provide filtered modulation symbols, wherein the first matrix is representative of an estimate of channel characteristics between a plurality of transmit antennas and the plurality of receive antennas used for the data transmission;

multiplying the filtered modulation symbols with a second matrix to provide estimates of transmitted modulation symbols; and

estimating characteristics of a plurality of transmission channels used for the data transmission.

19. The method of claim 18, further comprising:

demodulating the modulation symbol estimates in accordance with a particular demodulation scheme to provide demodulated symbols.

20. The method of claim 19, further comprising:

decoding the demodulated symbols in accordance with a particular decoding scheme.

21. The method of claim 18, further comprising:

combining modulation symbol estimates for redundant transmission to provide combined modulation symbol estimates.

22. The method of claim 18, further comprising:

deriving a channel coefficient matrix based on the received modulation symbols, and

wherein the first matrix is derived from the channel coefficient matrix.

23. The method of claim 22, wherein the channel coefficient matrix is derived based on received modulation symbols corresponding to pilot data.

24. The method of claim 18, wherein the second matrix is an inverse square matrix derived based on the first matrix.

25. The method of claim 1, wherein the CSI is derived at the receiver unit based on an unbiased minimum mean square error (UMMSE) processing.

26. The method of claim 25, wherein the UMMSE processing includes

processing the received signals to derive received modulation symbols;

multiplying the received modulation symbols with a first matrix **M** to provide estimates of transmitted modulation symbols; and

estimating characteristics of a plurality of transmission channels used for the data transmission based on the received modulation symbol, and

wherein the first matrix **M** is selected to minimize a mean square error between the modulation symbol estimates and transmitted modulation symbols.

27. The method of claim 26, further comprising:

multiplying the modulation symbol estimates with a second matrix to provide unbiased estimates of the transmitted modulation symbols, and

wherein the characteristics of the transmission channels are estimated based on the unbiased modulation symbol estimates.

28. The method of claim 27, further comprising:

deriving the first matrix **M** based on based on the unbiased modulation symbol estimates and to minimize the mean square error between the unbiased modulation symbol estimates and the transmitted modulation symbols.

29. The method of claim 1, wherein the MIMO system implements orthogonal frequency division modulation (OFDM).

30. The method of claim 29, wherein the processing at each of the receiver unit and transmitter unit is performed for each of a plurality of frequency subchannels.

31. A method for transmitting data from a transmitter unit to a receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

at the receiver unit,

receiving a plurality of signals via a plurality of receive antennas, wherein the received signal from each receive antenna comprises a combination of one or more signals transmitted from the transmitter unit,

processing the plurality of received signals to provide estimates of modulation symbols transmitted from the transmitter unit,

estimating signal-to-noise-plus-interference (SNR) of a plurality of transmission channels used for data transmission, and

transmitting SNR estimates for the transmission channels back to the transmitter unit; and

at the transmitter unit, processing data for transmission to the receiver unit in accordance with the received SNR estimates.

32. The method of claim 31, wherein the SNR of each of the plurality of transmission channels is estimated, and the SNR estimates for each transmission channel is transmitted back to the transmitter unit.

33. The method of claim 31, further comprising:

at the receiver unit,

deriving characterizations for the plurality of transmission channels used for data transmission, and

transmitting the characterizations back to the transmitter unit.

34. The method of claim 33, further comprising:

at the transmitter unit, pre-conditioning modulation symbols prior to transmission to the receiver unit in accordance with characterizations for the plurality of transmission channels.

35. The method of claim 31, wherein the received modulation symbols are processed in accordance with a channel correlation matrix inversion (CCMI) scheme.

36. The method of claim 31, wherein the received modulation symbols are processed in accordance with a minimum unbiased mean square error (UMMSE) scheme.

37. The method of claim 31, wherein the processing at the transmitter unit includes

coding data for each transmission channel in accordance with the received SNR estimate for the transmission channel.

38. The method of claim 37, wherein the processing of the processing at the transmitter unit further includes

modulating coded data for each transmission channel based on a modulation scheme selected based on the received SNR estimate for the transmission channel.

39. A multiple-input multiple-output (MIMO) communication system, comprising:

a receiver unit comprising

a plurality of front-end processors configured to receive a plurality of signals via a plurality of receive antennas and to process the received signals to provide received modulation symbols,

at least one receive MIMO processor coupled to the front-end processors and configured to receive and process the received modulation symbols to derive channel state information (CSI) indicative of characteristics of a plurality of transmission channels used for data transmission, and

a transmit data processor operatively coupled to the receive MIMO processor and configured to process the CSI for transmission back to the transmitter unit; and

a transmitter unit comprising

at least one demodulator configured to receive and process one or more signals from the receiver unit to recover the transmitted CSI, and

a transmit data processor configured to process data for transmission to the receiver unit based on the recovered CSI.

40. A receiver unit in a multiple-input multiple-output (MIMO) communication system, comprising:

a plurality of front-end processors configured to receive a plurality of transmitted signals via a plurality of receive antennas and to process the received signals to provide received modulation symbols;

a filter operatively coupled to the plurality of front-end processors and configured to filter the received modulation symbols in accordance with a first matrix to provide filtered modulation symbols, wherein the first matrix is representative of an estimate of channel characteristics between a plurality of transmit antennas and the plurality of receive antennas used for the data transmission;

a multiplier coupled to the filter and configured to multiply the filtered modulation symbols with a second matrix to provide estimates of transmitted modulation symbols;

a channel quality estimator coupled to the multiplier and configured to estimate characteristics of a plurality of transmission channels used for the data transmission and provide channel state information (CSI) indicative of the estimated channel characteristics; and

a transmit data processor configured to receive and process the CSI for transmission from the receiver unit.

41. The receiver unit of claim 40, further comprising:

a second estimator configured to derive a channel coefficient matrix based on the modulation symbol estimates, and wherein the first matrix is derived based on the channel coefficient matrix.

42. The receiver unit of claim 40, wherein the estimates of the transmission channel characteristics comprise signal-to-noise-plus-interference (SNR) estimates.

43. The receiver unit of claim 40, further comprising:

one or more demodulation elements, each demodulation element configured to receive and demodulate a respective stream of modulation symbol estimates in accordance with a particular demodulation scheme to provide a stream of demodulated symbols.

44. The receiver unit of claim 43, further comprising:

one or more decoders, each decoder configured to receive and decode a stream of demodulated symbols in accordance with a particular decoding scheme to provide decoded data.

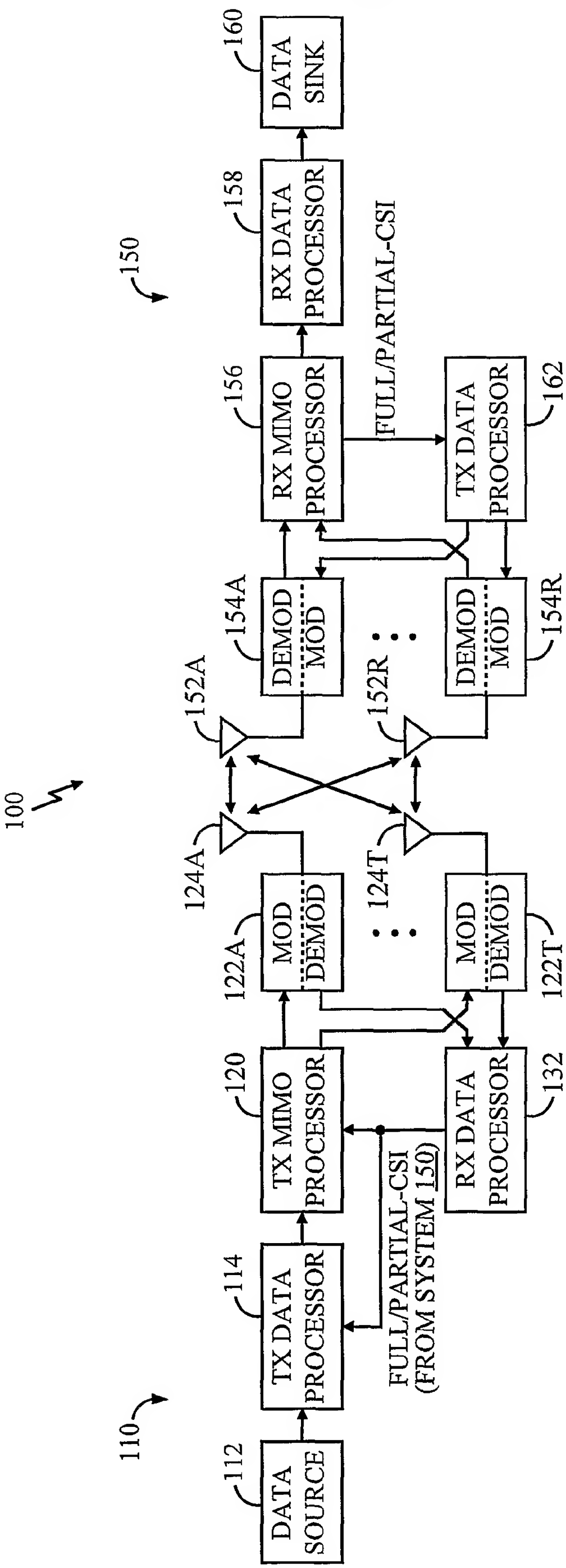

100
110
112
DATA SOURCE
114
TX DATA PROCESSOR
120
TX MIMO PROCESSOR
FULL/PARTIAL-CSI (FROM SYSTEM 150)
132
RX DATA PROCESSOR
122A
MOD
DEMOD
122T
MOD
DEMOD
124A
124T
152A
152R
154A
DEMOD
MOD
154R
DEMOD
MOD
150
156
RX MIMO PROCESSOR
158
RX DATA PROCESSOR
160
DATA SINK
FULL/PARTIAL-CSI
162
TX DATA PROCESSOR

FIG. 1

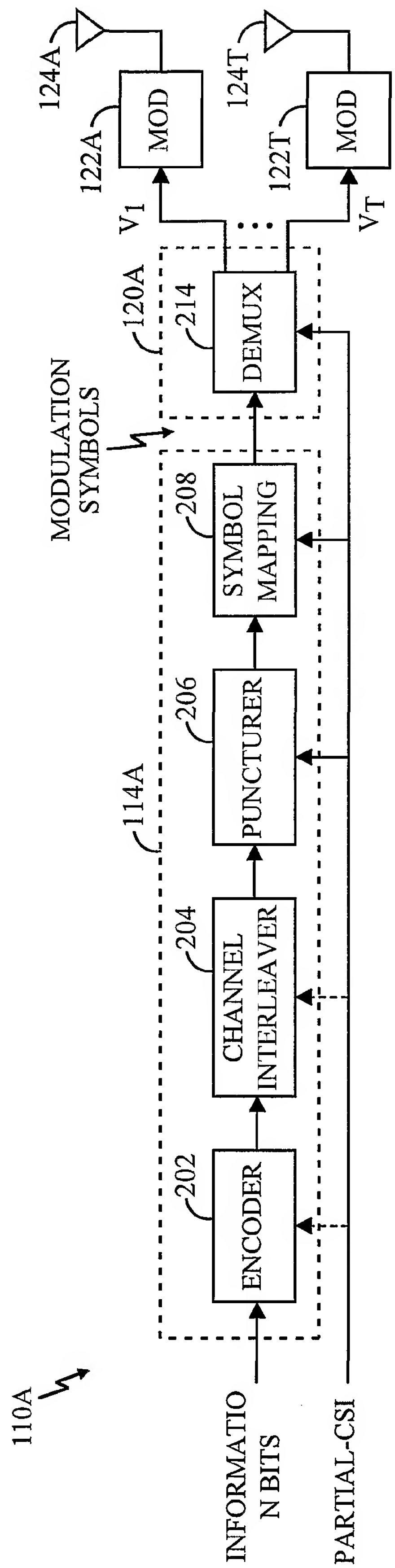

110A
INFORMATIO
N BITS
PARTIAL-CSI
114A
202
ENCODER
204
CHANNEL
INTERLEAVER
206
PUNCTURER
208
SYMBOL
MAPPING
MODULATION
SYMBOLS
120A
214
DEMUX
$V_1$
$V_T$
122A
MOD
122T
MOD
124A
124T

FIG. 2A

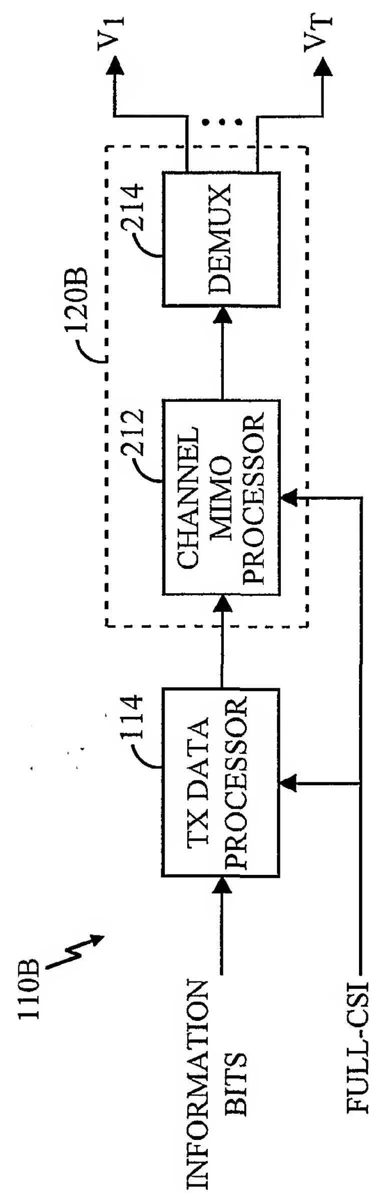

110B
INFORMATION
BITS
FULL-CSI
114
TX DATA
PROCESSOR
120B
212
CHANNEL
MIMO
PROCESSOR
214
DEMUX
$V_1$
$V_T$

FIG. 2B

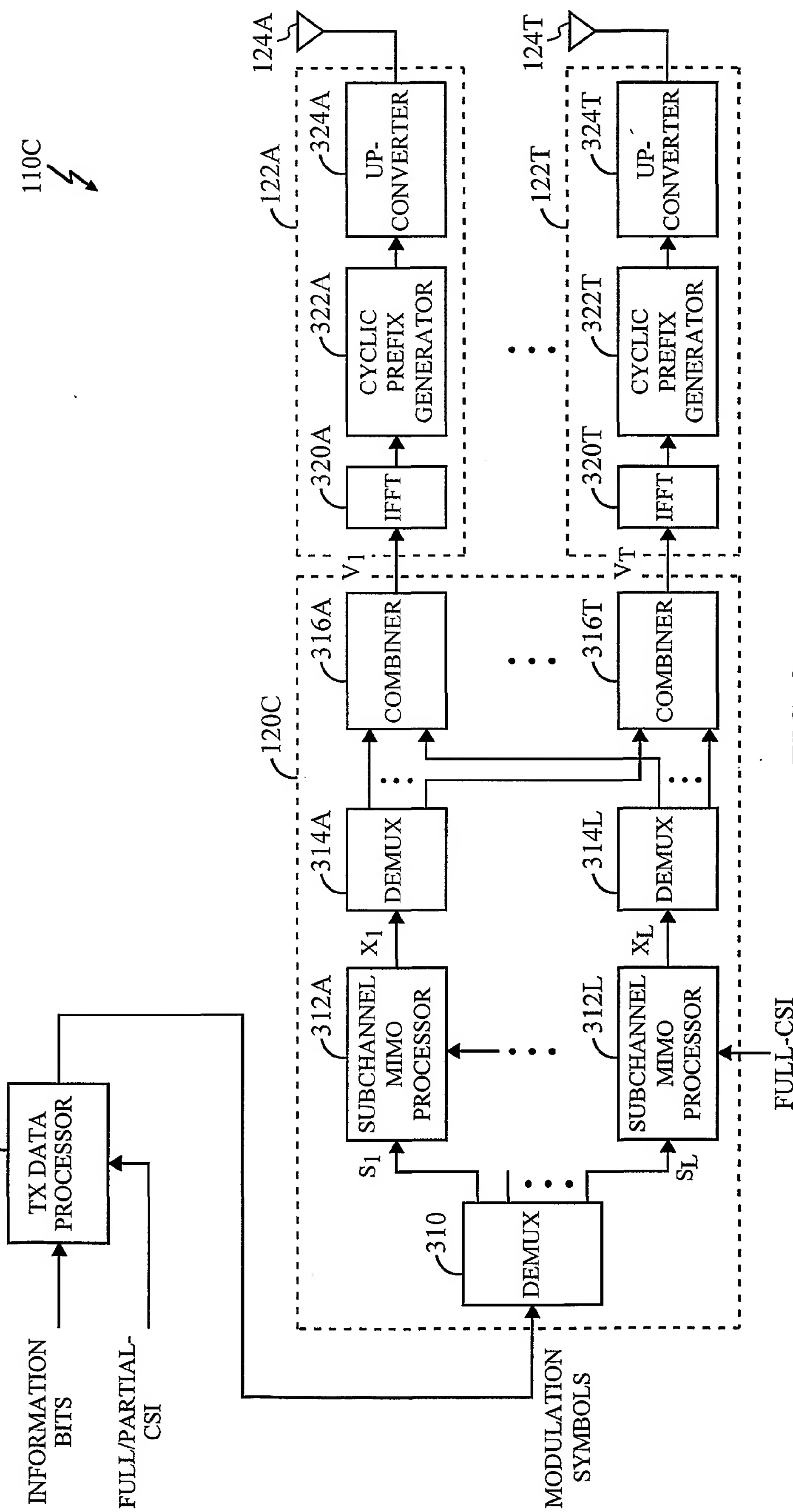

110C
INFORMATION BITS
FULL/PARTIAL-CSI
114
TX DATA PROCESSOR
MODULATION SYMBOLS
120C
310
DEMUX
$S_1$
$S_L$
312A
SUBCHANNEL MIMO PROCESSOR
312L
SUBCHANNEL MIMO PROCESSOR
FULL-CSI
$X_1$
$X_L$
314A
DEMUX
314L
DEMUX
316A
COMBINER
316T
COMBINER
$V_1$
$V_T$
122A
320A
IFFT
322A
CYCLIC PREFIX GENERATOR
324A
UP-CONVERTER
124A
122T
320T
IFFT
322T
CYCLIC PREFIX GENERATOR
324T
UP-CONVERTER
124T

FIG. 3

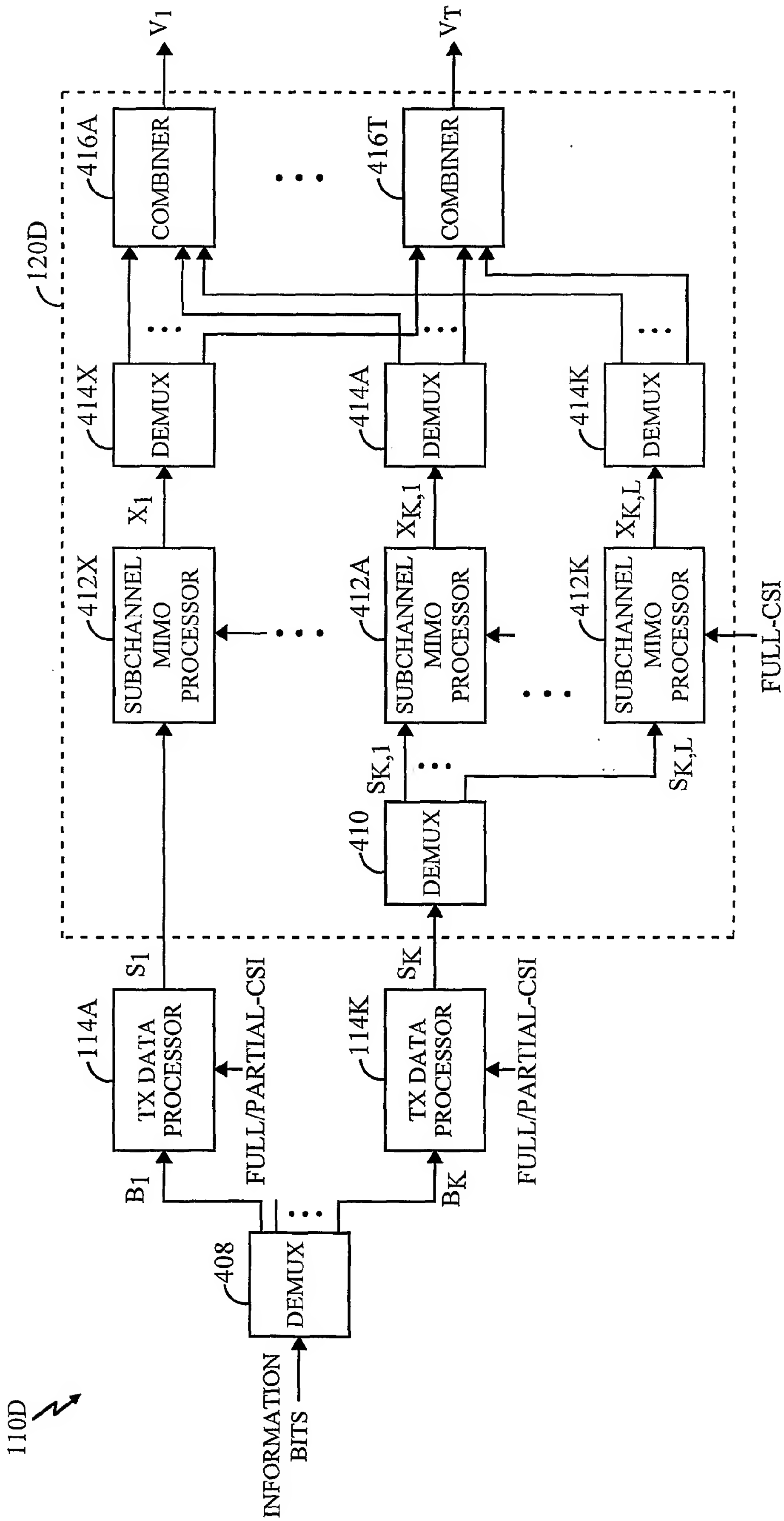

110D
INFORMATION BITS
408
DEMUX
B1
BK
114A
TX DATA PROCESSOR
FULL/PARTIAL-CSI
114K
TX DATA PROCESSOR
FULL/PARTIAL-CSI
S1
SK
120D
410
DEMUX
SK,1
SK,L
412X
SUBCHANNEL MIMO PROCESSOR
412A
SUBCHANNEL MIMO PROCESSOR
412K
SUBCHANNEL MIMO PROCESSOR
FULL-CSI
X1
XK,1
XK,L
414X
DEMUX
414A
DEMUX
414K
DEMUX
416A
COMBINER
416T
COMBINER
V1
VT

FIG. 4

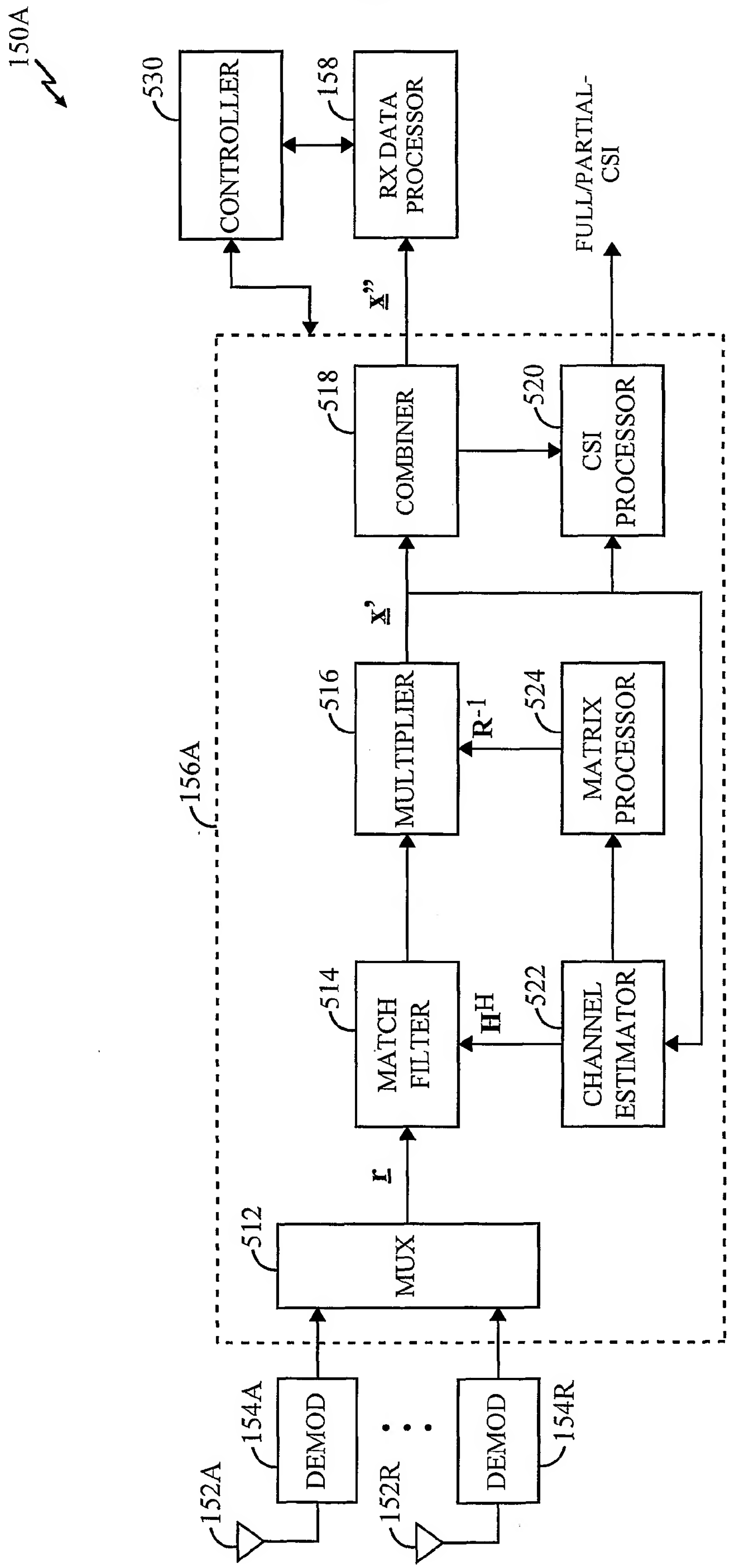

150A
530
CONTROLLER
158
RX DATA PROCESSOR
FULL/PARTIAL-CSI
x''
518
COMBINER
520
CSI PROCESSOR
x'
516
MULTIPLIER
R-1
524
MATRIX PROCESSOR
156A
514
MATCH FILTER
H^H
522
CHANNEL ESTIMATOR
r
512
MUX
152A
154A
DEMOD
152R
154R
DEMOD

FIG. 5

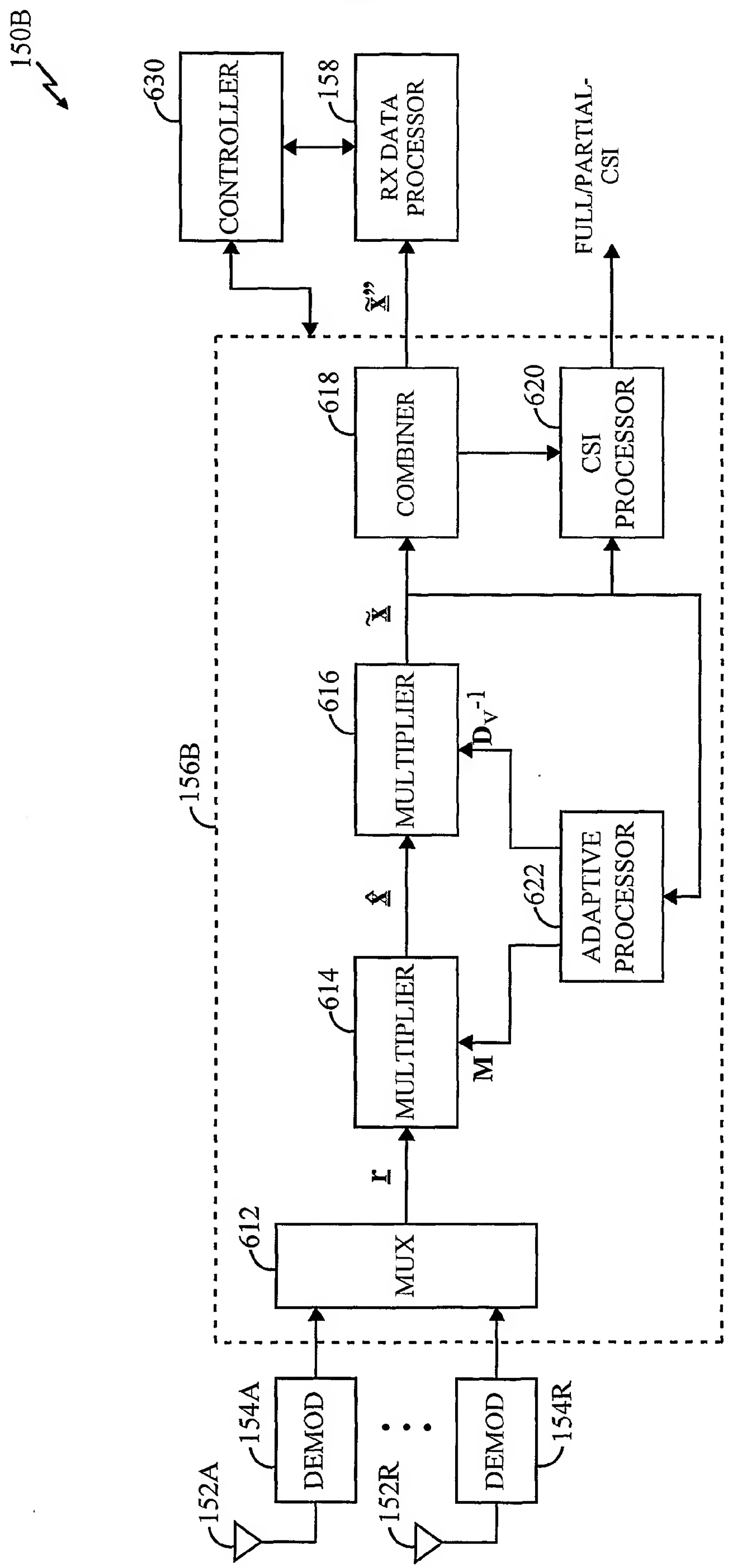

150B
CONTROLLER
630
RX DATA PROCESSOR
158
FULL/PARTIAL-CSI
156B
COMBINER
618
CSI PROCESSOR
620
MULTIPLIER
616
$\tilde{x}''$
$\tilde{x}$
$D_V^{-1}$
ADAPTIVE PROCESSOR
622
$\hat{x}$
MULTIPLIER
614
M
r
MUX
612
DEMOD
154A
152A
DEMOD
154R
152R

FIG. 6

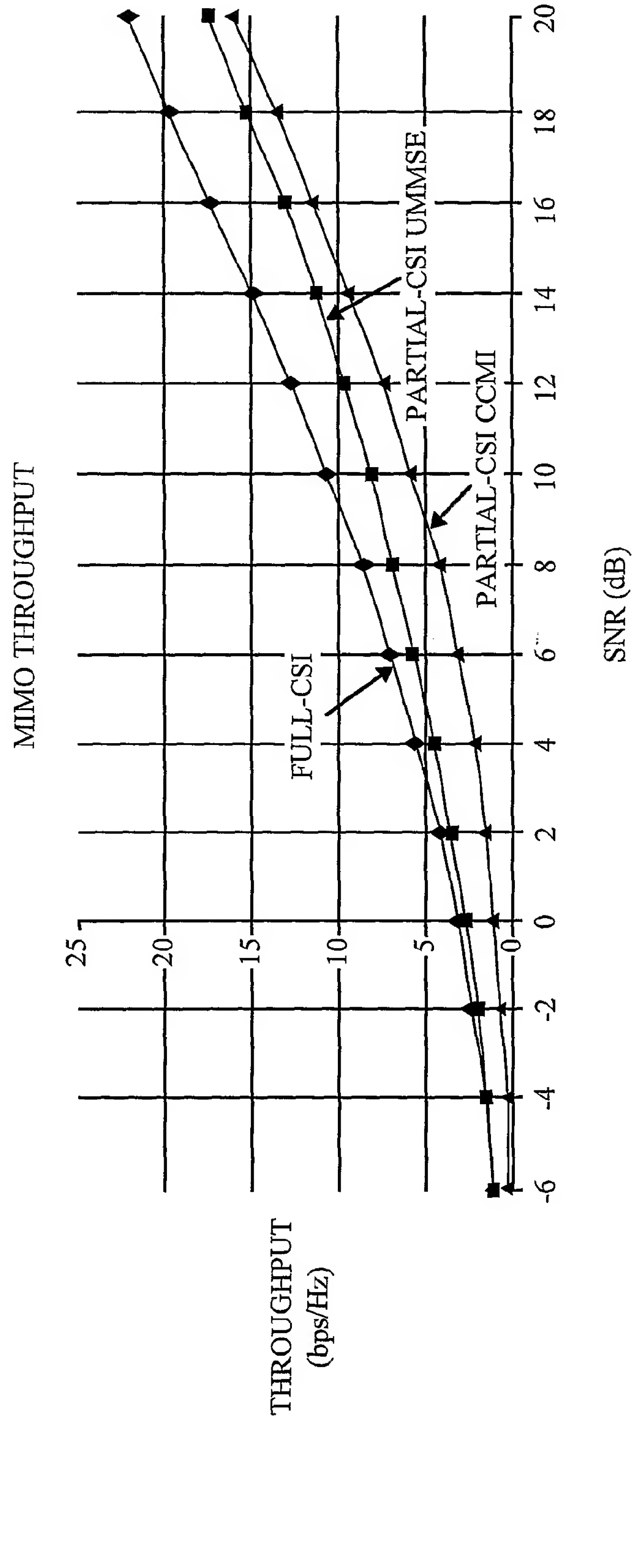

MIMO THROUGHPUT
THROUGHPUT
(bps/Hz)
25
20
15
10
5
0
-6
-4
-2
0
2
4
6
8
10
12
14
16
18
20
SNR (dB)
FULL-CSI
PARTIAL-CSI UMMSE
PARTIAL-CSI CCMI

FIG. 7A

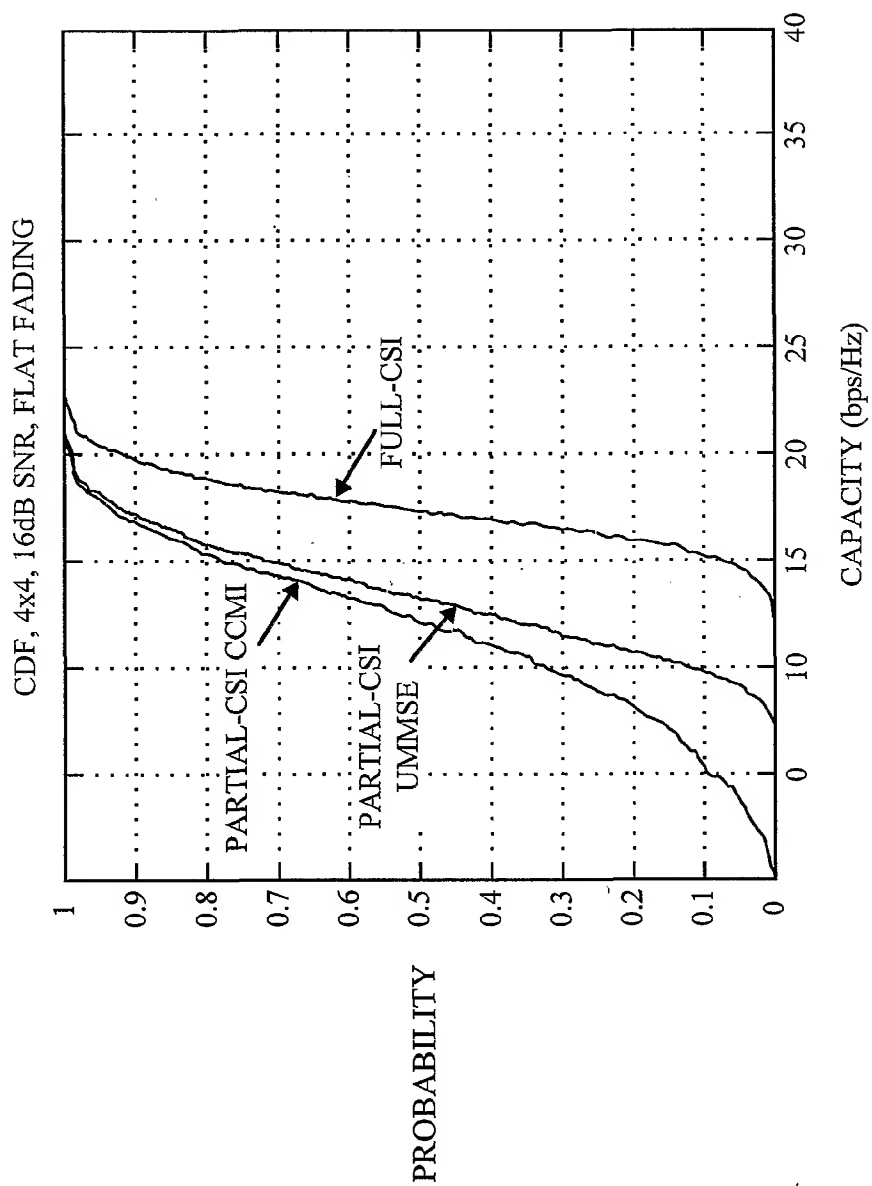

CDF, 4x4, 16dB SNR, FLAT FADING
PROBABILITY
CAPACITY (bps/Hz)
FULL-CSI
PARTIAL-CSI CCMI
PARTIAL-CSI UMMSE
0
0.1
0.2
0.3
0.4
0.5
0.6
0.7
0.8
0.9
1
0
10
15
20
25
30
35
40

FIG. 7B

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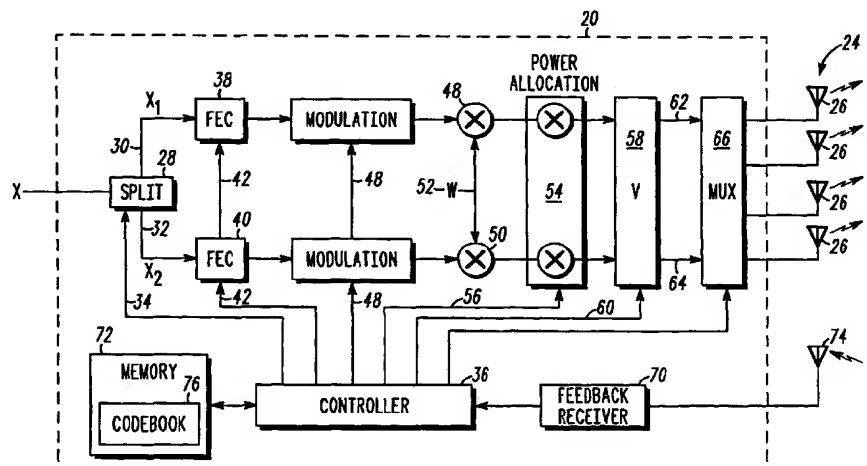
*[Continued on next page]*

(54) **Title:** METHOD AND SYSTEM IN A TRANSCEIVER FOR CONTROLLING A MULTIPLE-INPUT, MULTIPLE-OUTPUT COMMUNICATIONS CHANNEL




(57) **Abstract:** The present invention makes it possible to increase a data rate between a transmitter and receiver using a multiple-input, multiple-output radio frequency channel. A multiple-stream, multiple-antenna receiver measures a composite channel between a multiple-antenna transmitter and a multiple-antenna receiver to produce a composite channel measurement. The receiver selects a plurality of antenna array weight sets for use in the multiple-antenna transmitter in response to the composite channel measurement, where each antenna array weight set is associated with one of multiple data streams. Information describing the plurality of antenna array weight sets for use in the multiple-antenna transmitter are then transmitted.

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# METHOD AND SYSTEM IN A TRANSCEIVER FOR CONTROLLING A MULTIPLE-INPUT, MULTIPLE-OUTPUT COMMUNICATIONS CHANNEL

## Field of the Invention

The present invention is related in general to communication systems, and more particularly to a method and system for controlling the transmitting and receiving of multiple data streams in a multiple-input, multiple-output communications channel.

## Background of the Invention

Communication system designers are always looking for ways to increase the capacity of a communications channel between a transmitter and receiver. A communications channel may be defined as a system that transmits a sequence of symbols from one point to another. For example, a cellular communications system includes a channel for wirelessly transmitting a sequence of symbols that represent voice or data, back and forth between the telephone system and subscriber unit. An increase in the capacity of this channel means an increase in the rate of transmitting symbols. And when more symbols are transmitted in the same amount of time, voice can sound better, and it may take less time to transfer data files.

To increase the capacity of a wireless communications channel, antenna arrays have been used at the transmitter to better focus the transmitted energy at the receiver. An antenna array is a group of

spaced apart antennas that each transmit an antenna signal that has a specific gain and phase relationship with the other antenna signals. When the antennas work together transmitting the antenna signals, they produce an antenna pattern that is more focused on the receiver than a pattern produced by a single antenna. Note that the process of changing the gain and phase of a signal to produce antenna signals may be referred to as "weighting" the signal using a set of "antenna array weights."

Because antenna arrays may similarly be used at a receiver to improve signal quality, use of antenna arrays at both the transmitter and receiver has also been proposed to increase channel capacity. When multiple antennas are used at the transmitter and receiver, the wireless channel between them may be referred to as a multiple-input, multiple-output (MIMO) channel.

Fig. 1 shows a high-level schematic diagram of a communications channel, wherein a portion of the communications channel is wireless. As shown, x represents user data that will be wirelessly transmitted to the receiver. At the receiver, x is represented as an estimate of the data, $\hat{\mathbf{x}}$. User data x may be split to produce a vector that represents multiple data streams, $x_1$, $x_2, \ldots$

User data x is processed by matrix **V** to produce adaptive array antenna signals z. Each column of matrix **V** is a vector containing an antenna array weight set used to transmit one of the data streams $x_i$. Signals z are transmitted from antenna elements of the antenna array, through the air, and received at the receiver antenna array as received antenna signals r. The air interface between antenna signals z and received antenna signals r includes matrix H, which describes the effects of the air interface on signals z. The air interface is also described by the addition of noise n to signals z.

Received antenna signals r are processed in the receiver by matrix U' to produce the estimate of data, $\hat{\mathbf{x}}$.

With reference now to **Fig. 2**, there is depicted a two-input, two-output MIMO antenna array system. This MIMO system may be used to simultaneously transmit two different data streams, $x_1$ and $x_2$, to a single subscriber unit through a "composite channel" **H**, defined by the matrix

$$\mathbf{H} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix}$$

where $h_{ij}$ , $i=1,2,\ j=1,2$ are complex channel values. Note that the term "composite channel" as used herein refers to a complete measurement or description of a channel, wherein the effects of all combinations of transmit antennas and receive antennas are considered. The composite channel may be thought of as the aggregation of all channels between pairs of single antennas, defined by all pair-wise combinations of transmit and receive antennas.

When a flat Rayleigh fading channel is assumed, $h_{ij}$ are complex-valued Gaussian numbers with unity average power, $E[h_{ij}h_{ij}^*]=1$. The received (baseband) vector **r** (see **FIG. 1**) can be written as follows

$$\mathbf{r} = \mathbf{Hx} + \mathbf{n}$$

where $\mathbf{x} = [x_1 \quad x_2]^T$ is the vector of transmitted data streams, and **n** is a vector of noise samples, with additive white Gaussian noise with variance $\sigma_n^2$.

Note that in a noise free channel, both streams can be recovered perfectly if channel matrix **H** is full rank. That is, two equations and two unknowns can be solved to recover the unknowns $\mathbf{x}=[x_1 \quad x_2]^T$. When $\mathbf{x}=\mathbf{H}^{-1}\mathbf{r}$, both data streams can be recovered and link, or channel, capacity can be doubled. For example, a linear architecture may use zero forcing receivers to multiply the received vector **r**, with $\mathbf{H}^{-1}$. This works well with a high signal-to-noise ratio (SNR), but with a low SNR it boosts noise, which is not desirable.

In another linear receiver architecture, a Minimum Mean Square Error (MMSE) receiver may be used to minimize the average difference between detected data streams and the received signal.

While linear and non-linear receiver architectures can both be implemented to detect the multiple streams in noisy channels, in practical applications, noise in the channel will often require the use of non-linear receivers, which are more complicated and expensive to build. Examples of non-linear receivers with improved performance are Serial-Interference-Cancellation (SIC) receivers and a Maximum Likelihood (ML) receivers. Because of their complexity and cost, non-linear receivers should be avoided if possible.

**Theoretical MIMO Capacity:**

The capacity of a MIMO system may be shown with the following analysis. Suppose the Singular Value Decomposition (SVD) of the channel matrix **H** is given by

$$\mathbf{H}=\mathbf{USV}' \tag{1}$$

where **S** is a diagonal matrix composed of the singular values (i.e., the square-roots of eigenvalues of $\mathbf{H}'\mathbf{H}$ or $\mathbf{HH}'$), **U** is an orthogonal matrix

with column vectors equal to the eigenvectors of $\mathbf{HH}'$, $\mathbf{V}$ is an orthogonal matrix with columns equal to the eigenvectors of $\mathbf{H}'\mathbf{H}$, and the "$'$" operator is the complex conjugate transpose operation. As an example, consider the following composite channel matrix

$$\mathbf{H} = \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \tag{1.1}$$

The SVD of this composite channel is

$$\mathbf{H} = \mathbf{USV}' = \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} \\ 1/\sqrt{2} & -1/\sqrt{2} \end{bmatrix} \begin{bmatrix} \sqrt{2} & 0 \\ 0 & \sqrt{2} \end{bmatrix} \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \tag{1.2}$$

Referring to **Fig. 1**, the transmit vector is

$$\mathbf{z} = \mathbf{Vx} \tag{2}$$

Thus, the received vector is

$$\mathbf{r} = \mathbf{Hz} + \mathbf{n} \tag{3}$$

Replacing $\mathbf{H}$ and $\mathbf{z}$ with (1) and (2), we get

$$\mathbf{r} = \mathbf{USV}'\mathbf{Vx} + \mathbf{n} = \mathbf{USx} + \mathbf{n} \tag{4}$$

where, since $\mathbf{V}$ is an orthonormal matrix, $\mathbf{V}'\mathbf{V}$ is replaced with identity. Next, the received vector is pre-multiplied with $\mathbf{U}'$:

$$\begin{aligned} \hat{\mathbf{x}} &= \mathbf{U}'\mathbf{USx} + \mathbf{U}'\mathbf{n} \\ &= \mathbf{Sx} + \mathbf{e} \end{aligned} \tag{5}$$

Again, since $\mathbf{U}$ is an orthonormal matrix, $\mathbf{U}'\mathbf{U}$ is replaced with identity. Note that the new noise vector, $\mathbf{e}$, has the same covariance

matrix as **n**, because pre-multiplication with an orthonormal matrix does not alter the noise covariance.

If equation (5) is rewritten for the case of 2 transmit antennas, and 2 receive antennas it becomes:

$$\begin{bmatrix} \hat{\mathbf{x}}_1 \\ \hat{\mathbf{x}}_2 \end{bmatrix} = \begin{bmatrix} \sqrt{\lambda_1} & 0 \\ 0 & \sqrt{\lambda_2} \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} e_1 \\ e_2 \end{bmatrix} = \begin{bmatrix} \sqrt{\lambda_1} x_1 + e_1 \\ \sqrt{\lambda_2} x_2 + e_2 \end{bmatrix} \tag{6}$$

where $\lambda_i$ are the channel matrix eigenvalues.

The error free channel capacity based on the Shannon bound is well known, and is given by

$$\mathrm{C}_1 = \log_2(1+\rho) \quad \text{bits/symbol} \tag{7}$$

where $\rho$ is the channel SNR. From (5) and (6), note that the MIMO channel capacity based on the Shannon bound is the sum of the capacities per data stream:

$$C_{\mathrm{MIMO}} = \sum_{i=1}^{M} C_i = \sum_{i=1}^{M} \log_2\left(1 + \frac{\rho}{M}\lambda_i\right) \quad \text{bits/symbol} \tag{8}$$

where M is the minimum number of antennas at either the transmitter or the receiver. For the 2 transmit antenna, 2 receive antenna example, M=2. It is important to note that in (8), the total transmit power has been normalized such that it remains the same for any number of transmit antennas. The ratio $\rho / M$ ensures equal power transmitted on all antennas, and it maintains the same total power for all values of M.

In general, equal power transmission of (8) is sub-optimal. The total capacity, which is the sum of each data stream capacity, $C_{\text{MIMO}} = \sum_i C_i$, can be maximized by increasing the power to the high SNR streams, and reducing the power to the low SNR streams, such that the total transmit power remains the same. This procedure is typically referred to as "waterfilling."

By including waterfilling weights for optimum power allocation per data stream, (8) becomes:

$$C_{\text{MIMO}} = \sum_{i=1}^{M} \log_2\left(1 + \frac{\rho}{M}\lambda_i w_i\right) \quad \text{bits/symbol} \qquad (9)$$

where waterfilling weights are computed from

$$\sum_i w_i = \sum_i \max\left[0, \left(K - \frac{\sigma_n^2}{\lambda_i}\right)\right] = 1,$$

which is the waterfilling criterion, which is discussed by R.G. Gallager in *Information Theory and Reliable Communication,* New York: John Wiley & Sons, 1968. Here, $K$ is a constant determined by iterations, and $w_i$ are set accordingly.

Because transmitters in prior art systems lack data regarding the conditions of the composite channel, the performance of these systems cannot approach the Shannon bound for the MIMO channel. Furthermore, the amount of data needed to describe a composite MIMO channel is large, which would consume a large percentage of channel capacity when communicated to the transmitter.

Thus, it should be apparent that a need exists for an improved method and system for using feedback to efficiently control data

transmission and reception in a multiple-input, multiple-output radio frequency channel.

## Brief Description of the Drawings

The novel features believed characteristic of the invention are set forth in the appended claims. The invention itself, however, as well as a preferred mode of use, further objects, and advantages thereof, will best be understood by reference to the following detailed description of an illustrative embodiment when read in conjunction with the accompanying drawings, wherein:

**FIG. 1** is a high-level schematic diagram of a communications channel, wherein a portion of the communications channel is wireless;

**FIG. 2** is a high-level block diagram of a two-input, two-output MIMO channel;

**FIG. 3** is a block diagram of a multiple-stream, multiple-antenna transmitter that may be used to implement the method and system of the present invention;

**FIG. 4** is a more detailed block diagram of antenna array signal processor;

**FIG. 5** depicts a receiver for use in a multiple-stream, multiple-antenna transceiver system in accordance with the method and system of the present invention;

**FIG. 6** is a high-level logic flowchart that illustrates a feedback method in a multiple-stream, multiple-antenna receiver in accordance with the method and system of the present invention;

**FIG. 7** is a high-level logic flow chart that illustrates a feedback method in a multiple-stream, multiple-antenna transmitter in accordance with the method and system of the present invention;

**FIG. 8** is a more detailed logical flow chart that illustrates the process for estimating a composite channel and selecting array weight sets in accordance with the method and system of the present invention;

**FIG. 9** shows simulation results comparing quantized MIMO feedback with un-quantized, ideal MIMO feedback, in accordance with the method and system of the present invention; and

**FIG. 10** shows simulation results for a MIMO transceiver system in accordance with the method and system of the present invention.

## Detailed Description of the Invention

With reference now to **FIG. 3**, there is depicted a multiple-stream, multiple-antenna transmitter that may be used to implement the method and system of the present invention. As illustrated, transmitter **20** receives user data **22** and transmits user data **22** using antenna array **24**, which comprises antenna elements **26**.

User data **22** enters data splitter **28**, which separates the user data stream into a plurality of data streams, such as data stream **30** and data stream **32**. While two data streams are shown in **FIG. 3**, data splitter **28** may produce any number of data streams. Data splitter **28** splits data in proportion to control signal **34**, which is produced by controller **36**. For example, control signal **34** may specify a ratio of 2-to-1, wherein two bits are sent to data stream **30** and one bit is sent to data stream **32**. This splitting ratio may specify an equal number of bits on both streams, or all data bits are sent to one stream.

Data streams **30** and **32** output by data splitter **28** are input into error correction encoders **38** and **40**. These error correction encoders may be implemented with a convolutional encoder, a turbo encoder, a block encoder, or the like. The type of encoding, and the rate of encoding is controlled by control signal **42**, which is output by controller **36**. Note that control signal **42** may set error correction encoders **38** and **40** to the same error encoding schemes, or different encoding schemes.

Outputs of error correction encoders **38** and **40** are coupled to inputs of modulators **44** and **46**. Modulators **44** and **46** may be implemented with linear or non-linear modulation schemes, including

all varieties of modulators that modulate amplitude and phase, and combinations of amplitude and phase. Examples of modulators that may be used include Binary Phase Shift Keying modulators (BPSK), Quadrature Phase Shift Keying modulators (QPSK), M-ary phase shift keying modulators, M-ary quadrature amplitude modulators (MQAM), and the like.

Control signal **48** selects the type of modulation used in modulators **44** and **46**. Control signal **48** is produced by controller **36**. According to the present invention, the modulation schemes in the data streams may be the same, or different.

The output of modulators **44** and **46** are coupled to inputs of spreaders **48** and **50**, respectively. Spreaders **48** and **50** spread the signal using spreading code **52**, wherein the spreading code is assigned to user data **22**.

Outputs of spreaders **48** and **50** are coupled to inputs of power allocator **54**. Power allocator **54** sets a power ratio between data streams **30** and **32** in response to control signal **56** from controller **36**. Power allocator **54** may allocate all power to one data stream, equal powers on data streams, or other ratios of unequal power allocations.. Power allocator **54** does not allocate power to data streams **30** and **32** relative to data streams belonging to other user data not shown in **FIG. 3**. This means that power allocator **54** does not allocate an absolute level of power to a user. The absolute power allocated to each data stream, and each user, is determined by available power in power amplifiers and other control functions not shown in **FIG. 3**.

Outputs of power allocator **54** are coupled to inputs of antenna array signal processor **58**, which further processes the data streams

by applying antenna array weight sets to each data stream. These antenna array weight sets come from controller **36** via control signal **60**. By applying the antenna array weight sets to data streams **30** and **32**, antenna array signal processor enables the transmission of each data stream with a different antenna array pattern.

The outputs of antenna array signal processor **58** include weighted components of the input data streams. For example, output **62** may include a phase-and-gain weighted portion of data stream **30** added together with a phase-and-gain weighted portion of data stream **32**. The number of weighted outputs from antenna array signal processor **58** may be equal to or greater than the number of data streams. While the number of outputs of antenna array signal processor **58** may be greater than the number of data streams input, the number of data streams transmitted remains the same.

With reference now to **FIG. 4**, there is depicted a high-level block diagram of antenna array signal processor **58**. As shown, data streams **30** and **32** enter antenna array signal processor **58**, wherein a copy of each data stream is sent to a gain multiplier corresponding to an antenna element that will be used in an antenna array. In the example shown in **FIG. 4**, two antennas will be used in the antenna array, therefore copies of each data stream are sent to two gain multipliers **80**.

Following each gain multiplier **80** is a phase shifter **82**, which rotates the phase of the signal according to a control signal input. Outputs of phase shifters **82** are coupled to summers **84**, which add the weighted data streams to produce output signals **62** and **64**.

Control signal **60** (see **FIG. 3**) includes a plurality of antenna array weight sets, wherein one antenna array weight set is associated

with each data stream. For example, control signal **60** includes weight set signals **86** and **88**. Weight set signal **86** includes gain and phase weights (i.e., complex weights) for each gain multiplier **80** and phase shifter **82** associated with data stream **30**. Thus, the outputs of phase shifters **82** associated with data stream **30** produce antenna signals that provide a selected antenna pattern for data stream **30**. Similarly, weight set signal **88** includes phase and gain weights for each gain multiplier **80** and phase shifter **82** associated with data stream **32**. In the outputs of phase shifters **82** associated with data stream **32** produce antenna signals for driving an antenna array with a selected pattern for data stream **32**.

In order to produce desired antenna patterns for each data stream, gain multipliers **80** associated with a data stream may have different gain values and phase shifters **82** associated with a data stream may have different phase shift values, whereby producing antenna signals that work together to form a particular transmission pattern.

In some embodiments of transmitter **20**, output signals **62** and **64** may be up-converted, amplified, and coupled to two antenna elements **26**. However, in the embodiment shown in **FIG. 3**, multiplexer **66** is used to couple output signals **62** and **64** to selected antenna elements **26** in response to control signal **68** from controller **36**. This means that control signal **62** may be coupled to any one of antenna elements **26** in antenna array **24**, while output signal **64** is coupled to one of the remaining antenna elements **26**.

Controller **36** outputs control signals **34**, **42**, **48**, **56**, **60**, and **68** based upon information received from feedback receiver **70**, and data stored in memory **72**. Feedback receiver **70** is shown coupled to antenna **74** for receiving feedback data from a remote receiver, such

as the receiver shown in **FIG. 5**. While antenna **74** is shown separate from antenna array **24**, one of the antenna elements **26** of array **24** may be used to receive the feedback data.

Feedback data from feedback receiver **70** may include a codebook index, which may be used by controller **36** to lookup transmission parameters in codebook **76** within memory **72**.

Controller **36** may also be used to calculate, or derive, additional control signals or transmission parameters based upon feedback data. Therefore, it should be understood that feedback data may include measurements upon which calculations may be based, or data that indicates parameters to be used in transmitter **20**.

With reference now to **FIG. 5**, there is depicted a receiver for use in a multiple-stream, multiple-antenna transceiver system in accordance with the method and system of the present invention. As shown, receiver **98** includes antenna array **100** having elements **102** that receive radio frequency signals **104** and **106**. Received RF signals **104** and **106** are most likely different signals because antenna elements **102** are spaced apart, and propagation paths taken by received RF signals **104** and **106** from antenna elements **26** of transmitter **20** are most likely different in a multi-path fading environment.

In the multiple-stream, multiple-antenna transceiver system that is made up of transmitter **20** and receiver **98**, multiple data streams are transmitted to increase the data throughput between transmitter **20** and receiver **98**. Transmitter **20** is able to simultaneously transmit multiple data streams, and receiver **98** is able to keep the multiple streams separate by exploiting the differences in the channel characteristics between the multiple

antennas at transmitter **20** and receiver **98**. Thus, user data **22** in transmitter **20** is received by receiver **98** and output as estimated user data **108**.

Received RF signals **104** and **106** are input into radio frequency receiver front end **110**, wherein the radio frequency signals are down converted and digitized. The output of radio frequency receiver front end **110** is a stream of complex baseband digital samples that represent received RF signals **104** and **106**.

The outputs of radio frequency receiver front end **110** are input into receiver signal processor **112**, which has the function of separating data streams **30** and **32** (See **FIG. 3**) in receiver **98**. In one embodiment of the present invention, receiver signal processor **112** may be implemented by multiplying the input signals by the complex conjugate transpose of the **U** matrix, which is the left singular vectors of the singular value decomposition of the composite channel matrix **H**. Receiver signal processor **112** is controlled by control signal **115** from controller **113**.

The data streams output by receiver signal processor **112** are input to despreaders **114** and **116**, which despread the signals using spreading code **52**, which is the same spreading code used in the transmitter. The outputs of despreader **114** and **116** are coupled, respectively, to the inputs of demodulator and decoders **118** and **120**. Each demodulator and decoder **118** and **120** demodulates the signal and decodes the signal using demodulation and error correction decoding techniques that compliment those selected for each data stream in the transmitter. Thus, the type of demodulator and decoder functions used depends upon what was used in transmitter **20**, as indicated by control signal **122** from controller **113**. Demodulators

and decoders **118** and **120** may be the same function, or may be different functions.

The outputs from demodulator and decoder **118** and **120** are input into combiner **124**, which combines the multiple streams received back into a single stream for output as estimated user data **108**. Combiner **124** operates under the control of controller **113**, as directed by control signal **126**. Because the received data streams may have different data rates, and because one data stream may have a data rate equal to zero, combiner **124** must reconstruct the user data in accordance with the way data was originally split by data splitter **28** in transmitter **20** in **FIG. 3**.

In order to control the transmission of multiple data streams via multiple antennas at the transmitter, receiver **98** must measure the composite channel and send feedback data to the transmitter. As shown, outputs of radio frequency front end **110** are also coupled to composite channel estimator **128**, which uses pilot signals transmitted from each antenna element **26** in transmitter **20** to measure the composite channel between the multiple input antennas and multiple output antennas. The function of composite channel estimator **128**, and many of the other functional blocks in the data feedback portion of receiver **98**, are described in more completely in reference to **FIG. 8**, below.

The output of composite channel estimator **128**, which is represented by the **H** matrix, is input into **V** matrix computer and selector **130**. The "computing function" of block **130** computes **V**, which is a matrix describing desired antenna array weight sets to be used for each data stream in transmitter **20**. The desired antenna array weight sets are computed based upon the composite channel measurement.

The “selector function” of block **130** is a quantizing function that selects antenna array weight sets that most closely match the desire antenna array weight sets. By performing quantization, the amount of feedback data required to instruct transmitter **20** how to transmit over the MIMO channel may be reduced.

The selected antenna array weight sets output by computer and selector **130** are input into SNR computer and power allocator **132**, wherein a signal to noise ratio is computed for each data stream hypothetically transmitted using the selected antenna array weight sets. Based upon the SNR computations, the power allocation function of block **132** allocates power to each data stream, wherein the power is allocated to maximize the data throughput based upon a waterfilling algorithm. Once power has been allocated to each data stream, final SNR calculations may be performed using the selected power allocation.

Modulator and coder **134** receives information from SNR computer and power allocator **132** that it uses to select an encoding scheme and a modulation scheme to be used in transmitter **20**. Generally, higher order modulators are selected for data streams having high signal-to-noise ratios.

Feedback transmitter **136** receives information from the **V** matrix computer and selector **130**, SNR computer and power allocator **132**, and modulator and coder selector **134**. This data represents calculations and selections made in receiver **98** that will be used to control the transmission modes of transmitter **20**. In a preferred embodiment, feedback transmitter **136** analyzes the data and selects a codebook value associated transmitter parameters that most closely match the transmitter parameters represented by the input data. Therefore, feedback transmitter **136** may include codebook **138** for

producing a codebook value that is transmitted to transmitter **20** via antenna **140**. Although antenna **140** is shown separate from receive antenna array **100**, antenna **140** may be one of the antenna elements **102** in receive antenna array **100**. Data transmitted by feedback transmitter **136** is received in transmitter **20** by feedback receiver **70**.

With reference now to **FIG. 6**, there is depicted a high-level logic flowchart that illustrates a feedback method in a multiple-stream, multiple-antenna receiver in accordance with the method and system of the present invention. As illustrated, the process begins at block **300**, and thereafter passes to block **302** wherein the composite channel between the multiple-antenna transmitter and the multiple-antenna receiver is measured. This measurement results in the formation of the **H** matrix that is made up of complex channel values, representing gains and phases, as discussed above in reference to **FIG. 2**. The composite channel measurement is made by analyzing received antenna signals **r** (See **FIG. 1**) that include received pilot signals transmitted by each antenna at the transmitter.

Next, the process selects an antenna array weight set associated with each data stream in response to the composite channel measurement, as depicted at block **304**. Note that each simultaneously transmitted data stream has an associated set of weights that are used for each array antenna at the transmitter. Each antenna array weight set is used to produce an antenna pattern for the associated data stream.

In a preferred embodiment, selected antenna array weight sets are determined by calculating the right singular vectors of the SVD of composite channel matrix **H**. This process is more completely described with reference to **FIG. 8**. To reduce the amount of data needed to represent the antenna array weight sets, the desired weight

sets are compared to weight sets in a codebook, and one or more codebook weight sets having the closest distance are selected. The codebook indicator may represent a single antenna array weight set, or a combination of antenna array weight sets.

Note that if predefined combinations of antenna array weight sets are used, a first amount of information may be transmitted to describe a first antenna array weight set, and a second amount of information may be transmitted to describe a second antenna array weight set, wherein the second amount of information may be less than the first amount of information. Similarly, if a second antenna array weight set is restricted, or constrained, to a have a predefined relationship to a first antenna array weight set, the amount of information needed to describe the second set is less than that needed to describe the first.

Once selected, the antenna array weight sets are transmitted to the transmitter, and the transmitter uses the weights to produce selected antenna patterns for each data stream, as illustrated at block **306**. Because of the volume of data that may be needed to represent a complex weight for each antenna, for each data stream, it may be advantageous to use techniques that reduce the number of data bits transmitted from the receiver to the transmitter. As mentioned above, a codebook may be used to store several predefined antenna array weight sets. The number of antenna array weight sets available will determine the resolution of the quantizing process that takes an ideal set of weight sets and maps it to one of the available antenna array weight sets. Note that quantizing errors may become excessive if the number of available antenna array weight sets is too small.

As mentioned above, another way to reduce the amount of feedback data is to constrain the transmitter to transmitting antenna

patterns that have selected relationships with one another. For example, in a preferred embodiment, the antenna patterns at the transmitter may be constrained to be orthogonal to one another. Thus, by specifying a first antenna pattern, any remaining patterns at the base may be calculated, at least partially, according to the constraint relationships. Therefore, in a transmitter that transmits two data streams, if a first antenna pattern is specified, the antenna pattern for the second data stream may be derived, or calculated, so that the second pattern is constrained to be orthogonal to (or have low correlation with) the first.

**Details on V Quantization**

The simplest method of quantizing a matrix is to quantize each element of the matrix individually. Unfortunately, this method is inefficient and will require the greatest number of feedback bits for a desired performance. V may be quantized with two basic approaches: "block" and "incremental" quantization. In the first approach, all columns of V are quantized at once. In the second approach, columns of V are quantized incrementally.

**Block V Quantization**

Because the **V** matrix is orthonormal, it has some structure that can be exploited to reduce the amount of feedback. For the 2-antenna transmitter and 2-antenna receiver case, the **V** matrix can be written as

$$\mathbf{V} = \begin{bmatrix} v_{11} & v_{12} \\ v_{21} & v_{22} \end{bmatrix} = \begin{bmatrix} \cos\alpha & \sin\alpha \\ e^{j\theta}\cdot\sin\alpha & -e^{j\theta}\cdot\cos\alpha \end{bmatrix},$$

where

$$\alpha = \cos^{-1}(v_{11}),$$
$$\theta = \angle v_{21}.$$

The entire **V** matrix can be represented by two real parameters. Using this representation, there is a sign ambiguity in the second column vector that must be handled at the receiver. Fortunately, the transmission remains orthogonal and an MMSE receiver handles the sign ambiguity automatically. The parameters $\alpha \in \left[0, \frac{\pi}{2}\right]$ and $\theta \in [0, 2\pi]$ are uniformly quantized to a desired level. Figure 5 shows that quantizing **V** with 5 bits (3 for $\boldsymbol{\theta}$, 2 for $\boldsymbol{\alpha}$) and using an MMSE receiver is within 0.4 dB of the unquantized case.

In general, a codebook of **V** matrices can be created and indexed. A technique such as vector quantization can be used to generate the codebook and also to create an efficient mapping between **V** and the codebook. Parametric quantization as used in the 2x2 case can also be extended to larger **V** matrices.

**Incremental V Quantization**

In this approach, the columns of V are repeatedly drawn from a codebook of antenna array weights. (For example, one may use the TX AA codebooks from the 3GPP standard, release 99, or extensions of these codebooks.) The correlation properties of the columns of V are mirrored by selecting successive antenna array weight sets from increasingly smaller subsets of the codebook. As will be shown below, this constrained search reduces the amount of feedback data.

The column of V corresponding to the highest quality stream is selected first. This column is selected as the antenna array weight set that produces the maximum power at the receiver. The entire codebook is searched for this weight set.

Next, a second column of V is selected. A subset of the antenna array codebook may be found by searching for a codebook entries that have a correlation below a desired correlation threshold. The correlation threshold may be set to zero to select an orthogonal subset. Then, the antenna array weight set that produces maximum power at the receiver is selected from the low correlation subset of the codebook.

If there are three data streams, the third column of **V** is selected from a subset of the subset of codebook entries that was searched for the second column of **V**. The subset contains antenna array weight sets with low correlation against the subset searched for the second column. This process continues for all streams.

Since successive columns of **V** are searched from successively smaller subsets of the antenna array codebook, successive columns of **V** can be represented with fewer feedback bits. In a 4-element antenna array codebook with 64 entries, the antenna array weight set for the first column of **V** can be represented with log2(64) = 6 bits. By selecting an appropriate correlation threshold, the second column of **V**'s weight set can be represented with 4 bits, a third column with 2 bits, and the fourth column with 0 bits (only 1 antenna array weight set is possible, given the correlation threshold constraint and the choice of the other 3 antenna array weight sets.) Therefore, the entire **V** matrix can be quantized with 12 bits.

The size of codebook subsets may not be integer powers of two (since their size is determined by the correlation threshold), which means that the successively computed weight sets are not efficiently quantized using an integer number of bits to separately represent each weight set. In this case, alternate embodiments may jointly code the weight sets using vector quantization, or use variable length code words to reduce the number of bits required to represent the entire **V** matrix. Note that these alternate embodiments still draw the antenna array weight sets from subsets of a single codebook of antenna array weight sets, with the difference being the source coding used to reduce the number of bits required to represent the **V** matrix.

In addition to feeding back selected antenna array weight sets, the receiver may also feedback data that allows the transmitter to select a forward error correction coding scheme, a modulation scheme, a power allocation for each data stream, and a selection of antennas in the transmit antenna array.

As shown in block **308,** the process may select a data rate for each data stream in response to the composite channel measurement, the selected antenna array weight set, and SNR for each data stream. In a preferred embodiment, the SNR for each data stream is used to lookup a combination of encoding and modulation techniques according to calculated performance curves, and assuming equal power is available for both data stream. This lookup will provide an aggregated data throughput. This throughput value is compared to a second lookup assuming that all the power is used in the data stream having the highest signal to noise ratio. The second lookup gives a second data throughput, and the encoding and modulation scheme at the particular power setting is selected based upon the maximum throughput.

In a preferred embodiment, the codebook shown in table 1 below may be used in a system that sends four bits of feedback from the receiver to the transmitter in order to specify modulation and error encoding schemes for each data stream, and power allocation for each data stream. Note that antenna array weight sets are not included in the codebook of table 1.

| Configuration # | Modulator #1 | Code #1 | Modulator #2 | Code #2 | Power 1 | Power 2 |
|---|---|---|---|---|---|---|
| 1 | QPSK | R=1/2 | ---- | R=1/2 | 1 | 0 |
| 2 | QSPK | R=1/2 | QPSK | R=1/2 | 0.5 | 0.5 |
| 3 | 16 QAM | R=1/2 | ---- | R=1/2 | 1 | 0 |
| 4 | 16 QAM | R=1/2 | QPSK | R=1/2 | 0.5 | 0.5 |
| 5 | 16 QAM | R=1/2 | 16 QAM | R=1/2 | 0.5 | 0.5 |
| 6 | 64 QAM | R=1/2 | ---- | R=1/2 | 1 | 0 |
| 7 | 64 QAM | R=1/2 | QPSK | R=1/2 | 0.5 | 0.5 |
| 8 | 64 QAM | R=1/2 | 16 QAM | R=1/2 | 0.5 | 0.5 |
| 9 | 64 QAM | R=1/2 | 64 QAM | R=1/2 | 0.5 | 0.5 |
| 10 | 256 QAM | R=1/2 | ---- | R=1/2 | 1 | 0 |
| 11 | 256 QAM | R=1/2 | QPSK | R=1/2 | 0.5 | 0.5 |
| 12 | 256 QAM | R=1/2 | 16 QAM | R=1/2 | 0.5 | 0.5 |
| 13 | 256 QAM | R=1/2 | 64 QAM | R=1/2 | 0.5 | 0.5 |
| 14 | 256 QAM | R=1/2 | 256 QAM | R=1/2 | 0.5 | 0.5 |

**Table 1**

After the data rate is selected, the process transmits the selected data rate to the transmitter so the transmitter can select data encoding and modulation schemes for each data stream, as illustrated at block **310**. In a preferred embodiment of the invention, the receiver computes data rates, encoding schemes, modulation schemes, and power levels for each data stream, and transmits data that indicates these selections to the transmitter. In an alternative embodiment, the receiver may transmit measurements, or data based upon measurements, to the transmitter so that the transmitter may select a data rate, an encoding scheme, a modulation scheme, and a power allocation for each data stream.

Once the feedback data is transmitted from the receiver to the transmitter, the process ends, as depicted at block **312**. Although an end to the receiver feedback process is shown at block **312**, the process may iteratively continue in the receiver, beginning again at block **302** with new composite channel measurements.

With reference now to **FIG. 7**, there is depicted a high-level logic flow chart that illustrates a feedback method in a multiple-stream, multiple-antenna transmitter in accordance with the method and system of the present invention. As illustrated, the process begins at block **400**, and thereafter passes to block **402** wherein the process transmits a pilot signal on each antenna of the antenna array. Each pilot signal is distinguishable from the others. For example, different spreading codes may be used, or the same spreading code may be shifted in time relative to the other array antennas. These pilot signals provide a reference signal for the composite channel measurement.

Next, the process receives indications of a selected array weight set, with one set per data stream, as illustrated at block **404**. The indications of selected array weight sets may be data that describe a set of gains and phases for antenna signals for each antenna, with a set for each data stream in the transmitter. In a preferred embodiment, the selected array weight sets used for each data stream may be specified through the use of a codebook value received from the receiver, wherein the codebook value is used to lookup preselected sets of array weights.

Similarly, the process receives data that indicates data rates for each data stream, as depicted at block **406**. By indicating the data rate for each stream, the feedback data may also be indicating an encoding scheme, and a modulation scheme. The relationship

between data rates and encoding and modulation schemes exists because different encoding and modulation schemes have different capacities. Therefore, the selection of a data rate may force the selection of particular encoding and modulation schemes.

Next, the process receives an indication of power allocation for each data stream, as illustrated at block **408**. Note that a codebook value may be used as the "indicator" that indicates data rates and power allocation for each data stream. As discussed above, a single codebook value may be used to specify an encoding scheme, a modulation scheme, and a power allocation. In some embodiments, specifying a data rate alone may specify the encoder, modulator, and power allocation. For example, if the data rate selected was zero, no power is allocated and the encoding and modulation schemes are irrelevant.

After receiving the feedback data, the process selects power settings, and encoding and modulation schemes for each data stream, as depicted at block **410**. In this step, these parameters may be selected according to a codebook value received. In alternative embodiments, some of these parameters may be calculated or derived from the feedback data received. For example, if the antenna pattern of the first data stream is indicated, the process in the transmitter may derive or calculate an antenna pattern used for the second data stream. This may be done when, for example, the second stream is constrained to be orthogonal to the first stream.

Once transmit parameters are selected as shown in block **410**, the process separates input data into data streams according to selected data rates supported by encoding and modulation schemes selected for each data stream, as depicted at block **412**. This process is implemented in data splitting function **28** shown in **FIG. 3**. As an

example, if data stream 1 operates at twice the rate of data stream 2, then two symbols are sent to data stream 1 and a single symbol is sent to data stream 2. Similarly, if one data stream has zero power allocated, all the data symbols are sent to the remaining data streams having some power allocated.

Next, the process encodes each data stream, as illustrated at block **414**. The process of encoding may be implemented with a block coder, a convolutional coder, a turbo coder, and the like.

After encoding, each data stream is modulated, as depicted at block **416**. This modulation may be implemented using a BPSK modulator, a QPSK modulator, a M-PSK modulator, a M-QUAM modulator (where M is the number of constellation points), and the like.

Following the modulating step, the process modifies the gain and phase of each modulated data stream according to respective selected array weight sets to produce data stream antenna signals for each array antenna, as illustrated at block **418**. Examples of data stream antenna signals are the outputs of phase shifters **82** in **Fig. 4**. The number of data stream antenna signals produced in this step equals the number of data streams times the number of antenna elements in the antenna array.

After producing data stream antenna signals for each array antenna, the data stream antenna signals associated with the same array antenna are summed to produce antenna signals, as depicted at block **420**. Examples of antenna signals are the outputs of summers **84** in **Fig. 4**. These antenna signals are combinations of signals from each data stream that have been weighted in gain and phase according to the selected array weight sets. This complex combination

of signals is more concisely described according to the **V** matrix used in the transmitter, which is discussed above in relation to **FIG. 1**.

Finally, the antenna signals for each antenna are transmitted, as illustrated at block **420**. The transmission step includes further processing, upconversion, and amplification needed for radio frequency transmission.

The feedback method ends, as depicted at block **424**. Although the process is shown with an end, the process may iteratively repeat in the transmitter in order to update each antenna pattern for each data stream in response to varying channel conditions.

Turning now to **FIG. 8**, there is depicted a more detailed logical flow chart that illustrates the process for estimating a composite channel and selecting array weight sets, which is shown at a higher level in **FIG. 6**. As illustrated, the process begins at block **500**, and thereafter passes to block **502** wherein the process estimates channel matrix **H** using received pilot signals, wherein a pilot signal is transmitted from each transmitter antenna. Pilots may or may not be orthogonal, but they are selected so that they are distinguishable at the receiver.

Next, the process computes a singular valve decomposition of matrix **H** to find matrix **V**, wherein $\mathbf{H} = \mathbf{U}\mathbf{S}\hat{\mathbf{V}}^{\mathrm{T}}$, as depicted at block **504**. Transmitting with this **V** matrix allows operation of the MIMO channel at near Shannon capacity for MIMO.

Thereafter, the process selects an index for a quantized **V** matrix, as illustrated at block **506**. The quantizing may be preformed by a codebook lookup, or other methods, discussed above. Note that

the quantized **V** matrix represents selected antenna array weight sets. Antenna array weight sets may be quantized as a group, or separately.

After quantizing, the process estimates a signal-to-noise ratio (SNR) of each data stream based on the transmitter using the quantized **V** matrix, and assuming equal power streams, as depicted at block **508**.

Next, the process uses the estimated SNR to determine power allocation of each data stream using a waterfilling algorithm, as illustrated at block **510**. An alternative to waterfilling is a brute-force search of all quantized possibilities. In a preferred embodiment, this parameter can be quantized to a low number of bits. For example, a reasonable choice for power allocation may be one-bit indicator for both streams "on", or only one stream "on" and the other "off".

Based on the power allocation for each stream, and the estimated SNR for each stream, the process next selects the coding method and modulation method, as depicted at block **512**. This may be implemented with a lookup that maps every SNR range to a modulator-encoder combination. In general, the coding and modulating is adapted for each data stream according to the channel quality. For example, if high channel quality is indicated by a high SNR, the modulator may be set to 16-QAM; otherwise, QPSK modulation may be selected.

Finally, the process transmits, to the transmitter, indicators for a quantized **V** matrix, a power allocation for each stream, and coding and modulation methods, as illustrated at block **514**. In a preferred embodiment, the process uses a codebook to indicate quantized antenna array weight sets, and other modulation parameters.

As depicted, the process ends at block **516**.

Referring again to **FIG. 3**, the number of antennas used by transmitter **20** is equal to the number of outputs from antenna array signal processor **58**. As shown in **FIG. 3**, antenna array signal processor **58** has two outputs, output signals **62** and **64**.

As mentioned earlier, output signals **62** and **64** may be transmitted from two antennas, or multiplexer **66** may be used to select two antennas to form an antenna array from a larger number of "available antennas", such as the four antenna elements **26** shown in antenna array **24**. Thus, in some embodiments of the present invention, there exists a set of available antenna elements, from which a subset of the "available antenna elements," from which a subset of the available antenna elements may be selected to form "an antenna array", wherein the antenna array comprises antenna elements actually used to transmit the multiple data streams.

While the embodiment in **FIG. 3** shows multiplexer **66** for selecting antennas, alternative embodiments may use the **V** matrix to select antennas mathematically by multiplying signals by zero, or non-zero values according to the matrix elements.

In order to select the antenna elements from the set of available antenna elements, receiver **98** measures a composite channel that includes all channels between all pair-wise selections of all available antenna elements and all antenna elements at the receiver. Thus, in **FIGS. 3** and **5**, between transmitter **20** with 4 available antennas and receiver **98** with 2 receive antennas, the composite channel measurement forms a composite channel matrix **H** that is four rows by two columns.

At the transmitter, there are 6 ways to choose 2 antennas from a set of 4 available antennas. The antenna array is formed with the pair that yields the highest capacity composite channel. The selection process may be described by the following expression:

$$\max_i \det\left(\mathbf{I} + \frac{1}{2\sigma^2}\mathbf{H}_i'\mathbf{H}_i\right) \tag{10}$$

where (without using waterfilling) half power is allocated to each data stream, $\sigma^2$ is the noise variance, $\mathbf{I}$ is the 2x2 identity matrix, and

$$\mathbf{H}_1 = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix}, \mathbf{H}_2 = \begin{bmatrix} h_{11} & h_{13} \\ h_{21} & h_{23} \end{bmatrix}, \mathbf{H}_3 = \begin{bmatrix} h_{11} & h_{14} \\ h_{21} & h_{24} \end{bmatrix},$$
$$\mathbf{H}_4 = \begin{bmatrix} h_{12} & h_{13} \\ h_{22} & h_{23} \end{bmatrix}, \mathbf{H}_5 = \begin{bmatrix} h_{12} & h_{14} \\ h_{22} & h_{24} \end{bmatrix}, \mathbf{H}_6 = \begin{bmatrix} h_{13} & h_{14} \\ h_{23} & h_{24} \end{bmatrix}.$$

To select one of the six pairs, three feedback bits are required. In order to reduce the feedback data even more, two bits can be used to select one of four pairs.

The receiver next considers all two-by-two combinations of transmit and receive antennas, wherein there are six possible combinations of two transmit antennas and two receive antennas. For each of the six combinations, an aggregate data rate is computed, wherein the aggregate data rate is the total data rate provided by adding the data rate of data stream 1 and the data rate of data stream 2. By ranking the aggregate data rates, the antenna combination that supports the highest data rate may be selected.

In an alternative embodiment of transmitter **20**, antenna array signal processor **58** may use a **V** matrix that produces four outputs to drive four antennas in an antenna array. However, the amount of feedback data necessary to support selection of antenna array weight

sets for a four-output **V** matrix begins to consume an unacceptable percentage of capacity of the link used for feedback data. Therefore, a two-output **V** matrix is used to drive two antennas that are selected from an available set of four antennas. The two antennas that are selected support the highest aggregate data rate between transmitter **20** and receiver **98**. In the transmitter that selects antenna elements from a larger set of available antenna elements a trade-off has been made between reducing uplink feedback data and reducing downlink performance.

It should be appreciated from the discussion above that the present invention makes it possible to increase a data rate between a transmitter and receiver using a multiple-input multiple-output radio frequency channel. The feedback method disclosed is a practical solution to controlling a MIMO transceiver.

Advantages of using the MIMO radio frequency channel include the ability to double an effective data throughput without using additional communication resources, such as spreading codes, power, and bandwidth, and without employing higher order modulators. In other words, using the same communication resources, with the same modulator, the throughput can be doubled by effectively controlling the MIMO radio frequency channel. This effective control of the channel involves transmitting multiple data streams in a way that they can be separated from one another at the receiver. This MIMO channel control exploits specific knowledge of the channel gained by measuring a composite channel between the transmitter and receiver. Furthermore, proper control of the MIMO channel enables the use of linear receivers, rather than the more complex or expensive non-linear receiver. By transmitting the signal vector $\mathbf{x}$ along the channel eigenmodes (i.e., transmitting $\mathbf{z}=\mathbf{V}\mathbf{x}$ rather than $\mathbf{x}$), we can completely separate the two streams without using non-linear detectors. Thus,

with the proper control of the MIMO channel, the non-linear receiver has no substantial advantage over the linear receiver.

**Fig. 9** shows simulation results comparing quantized MIMO feedback with un-quantized, ideal MIMO feedback. There is little degradation due to quantizing.

**Fig. 10** shows simulation results for a MIMO transceiver system described above. The codebook used for this simulation is found in Table 1. The **V** matrix is selected with 5 feedback bits, and the encoding, modulation, and power allocation are selected with 4 feedback bits. The simulation results show that a MIMO system with 9 bits of feedback performs about 4 dB from the theoretical MIMO Shannon bound. Note that if some combinations of modulator, coder, and power allocation occur infrequently, they can be removed with a small loss in performance, which further reduces the feedback bits needed.

The foregoing description of a preferred embodiment of the invention has been presented for the purpose of illustration and description. It is not intended to be exhaustive or to limit the invention to the precise form disclosed. Obvious modifications or variations are possible in light of the above teachings. The embodiment was chosen and described to provide the best illustration of the principles of the invention and its practical application, and to enable one of ordinary skill in the art to utilize the invention in various embodiments and with various modifications as are suited to the particular use contemplated. All such modifications and variations are within the scope of the invention as determined by the appended claims when interpreted in accordance with the breadth to which they are fairly, legally, and equitably entitled.

## Claims

What is claimed is:

**1.** A feedback method in a multiple-stream, multiple-antenna receiver, the method comprising the steps of:

measuring a composite channel between a multiple-antenna transmitter and a multiple-antenna receiver to produce a composite channel measurement;

selecting a plurality of antenna array weight sets for use in the multiple-antenna transmitter in response to the composite channel measurement, wherein each antenna array weight set is associated with one of multiple data streams; and

transmitting information describing the plurality of antenna array weight sets for use in the multiple-antenna transmitter.

**2.** The feedback method of claim **1,** wherein the step of selecting a plurality of antenna array weight sets further includes selecting a plurality of antenna array weight sets having a cross correlation less than the inverse of a number of antenna elements in the antenna array of the multiple-antenna transmitter.

**3.** The feedback method of claim **1** further including the steps of:

selecting a data rate for each data stream in response to the composite channel measurement; and

transmitting information describing the data rate selection for use in the multiple-antenna transmitter.

**4.** The feedback method of claim **1** further including the steps of transmitting information used to describe a quality of each data stream for use in the multi-antenna transmitter.

**5.** The feedback method of claim **1** wherein the step of selecting the plurality of antenna array weight sets further includes the steps of:

selecting a first antenna array weight set from a codebook having a plurality of preselected antenna array weight sets; and

selecting a second antenna array weight set from a subset of the codebook.

**6.** The feedback method of claim **1** further including the steps of:

measuring a composite channel between a multiple-antenna transmitter and a multiple-antenna receiver to produce a composite channel measurement, wherein pilot signals are received from M number of available antennas at the multiple-antenna transmitter;

selecting N antennas to be used at the transmitter, from M number of available antennas, in response to the composite channel measurement, wherein the N selected antennas will be used to form the antenna array at the multiple-antenna transmitter.

**7.** A feedback method in a multiple-stream, multiple-antenna transmitter, the method comprising the steps of:

splitting user data to produce multiple data streams;

transmitting a pilot signal from each antenna of an antenna array;

receiving indications of a selected antenna array weight set for each of the multiple data streams, wherein each antenna array weight set includes weights associated with each antenna of the antenna array;

using the selected antenna array weight sets, weighting each data stream to produce antenna signals for each antenna in the antenna array; and

transmitting the antenna signals, wherein the multiple data streams are transmitted.

**8.** The feedback method of claim **7** further including the steps of:

encoding and modulating each of multiple data stream to produce modulated data streams; and

using the selected antenna array weight sets, weighting each modulated data stream to produce antenna signals for each antenna in the antenna array.

**9.** The feedback method of claim **7** further including the steps of:

receiving indications of a selected data rate for each data stream;

splitting data in proportion to the selected data rates for each data stream; and

transmitting multiple data streams using the selected data rates for each data stream.

**10.** The feedback method of claim **9** further including the steps of:

selecting encoding and modulation schemes for each data stream in response to the selected data rate; and

transmitting multiple data streams using the selected encoding and modulation schemes for each data stream.

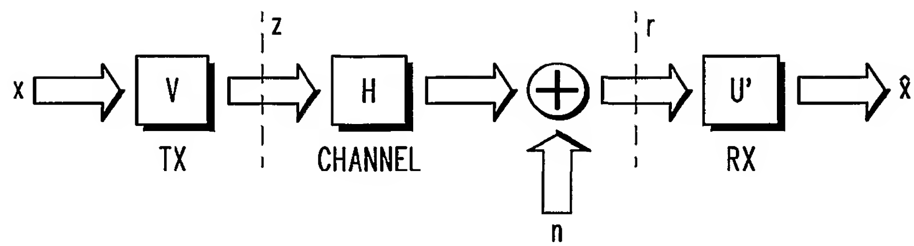

x
V
TX
z
H
CHANNEL
n
r
U'
RX
x̂

FIG. 1

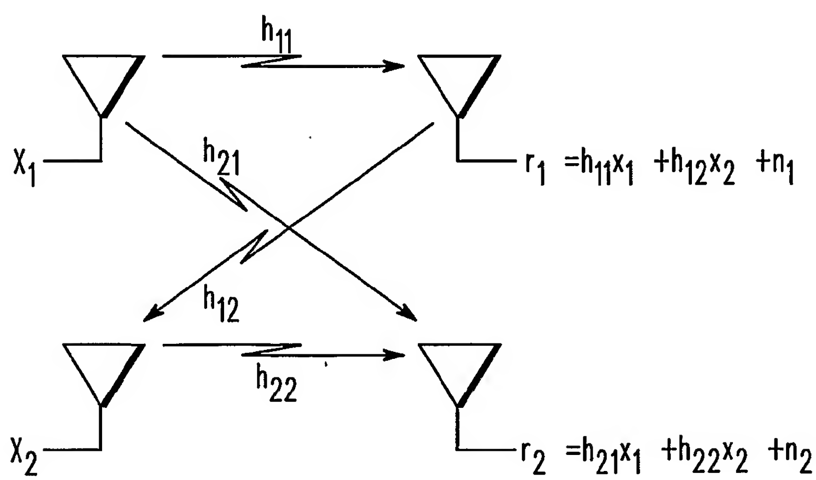

h11
h21
h12
h22
X1
X2
r1 =h11x1 +h12x2 +n1
r2 =h21x1 +h22x2 +n2

FIG. 2

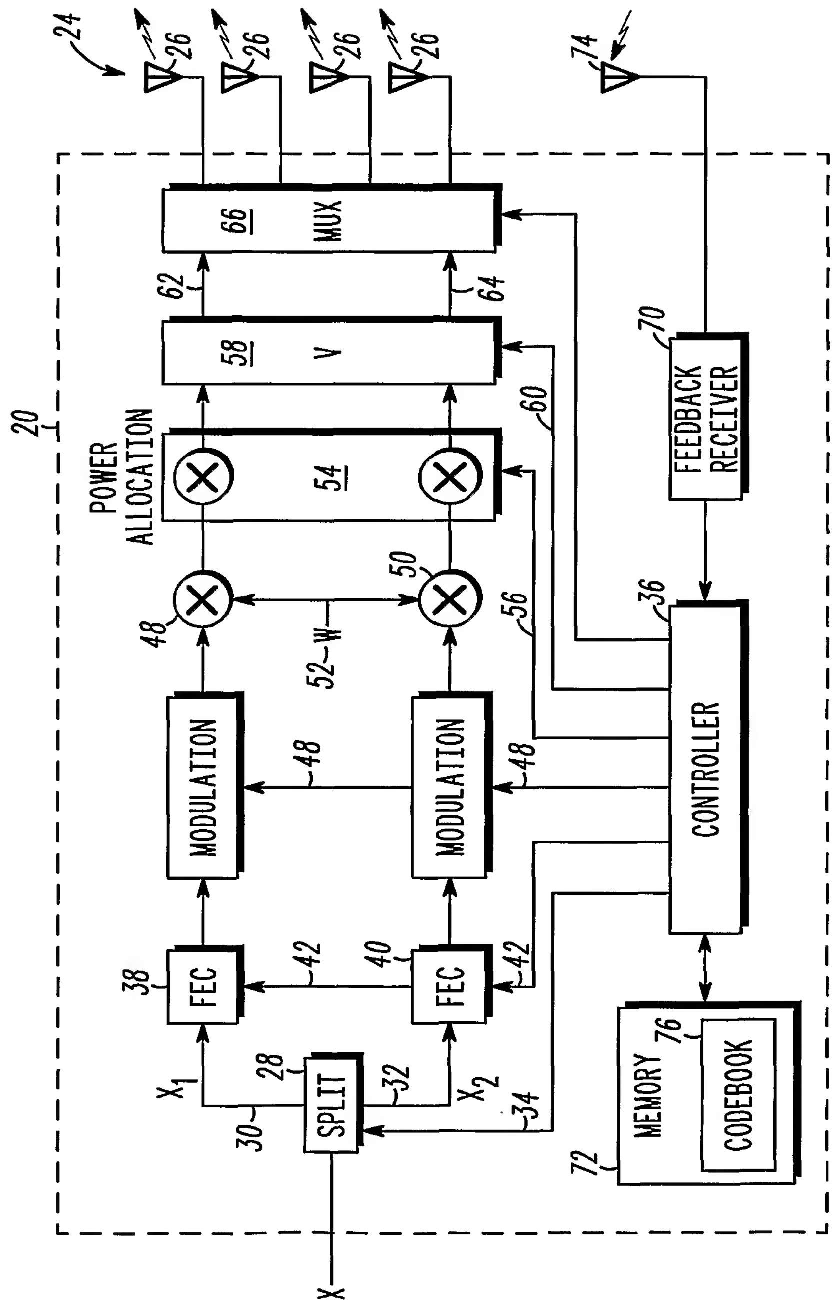

20
24
26
74
X
SPLIT
28
30
32
34
X1
X2
FEC
38
40
42
MODULATION
48
52~W
50
56
POWER ALLOCATION
54
58
V
60
62
64
66
MUX
CONTROLLER
36
FEEDBACK RECEIVER
70
MEMORY
72
CODEBOOK
76

FIG. 3

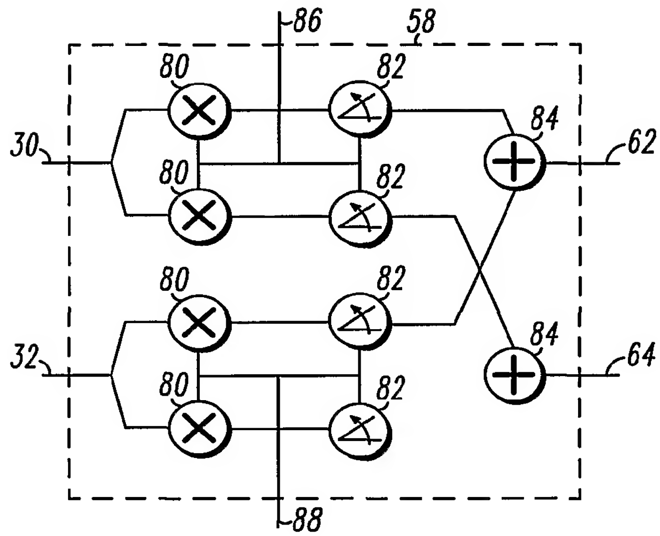
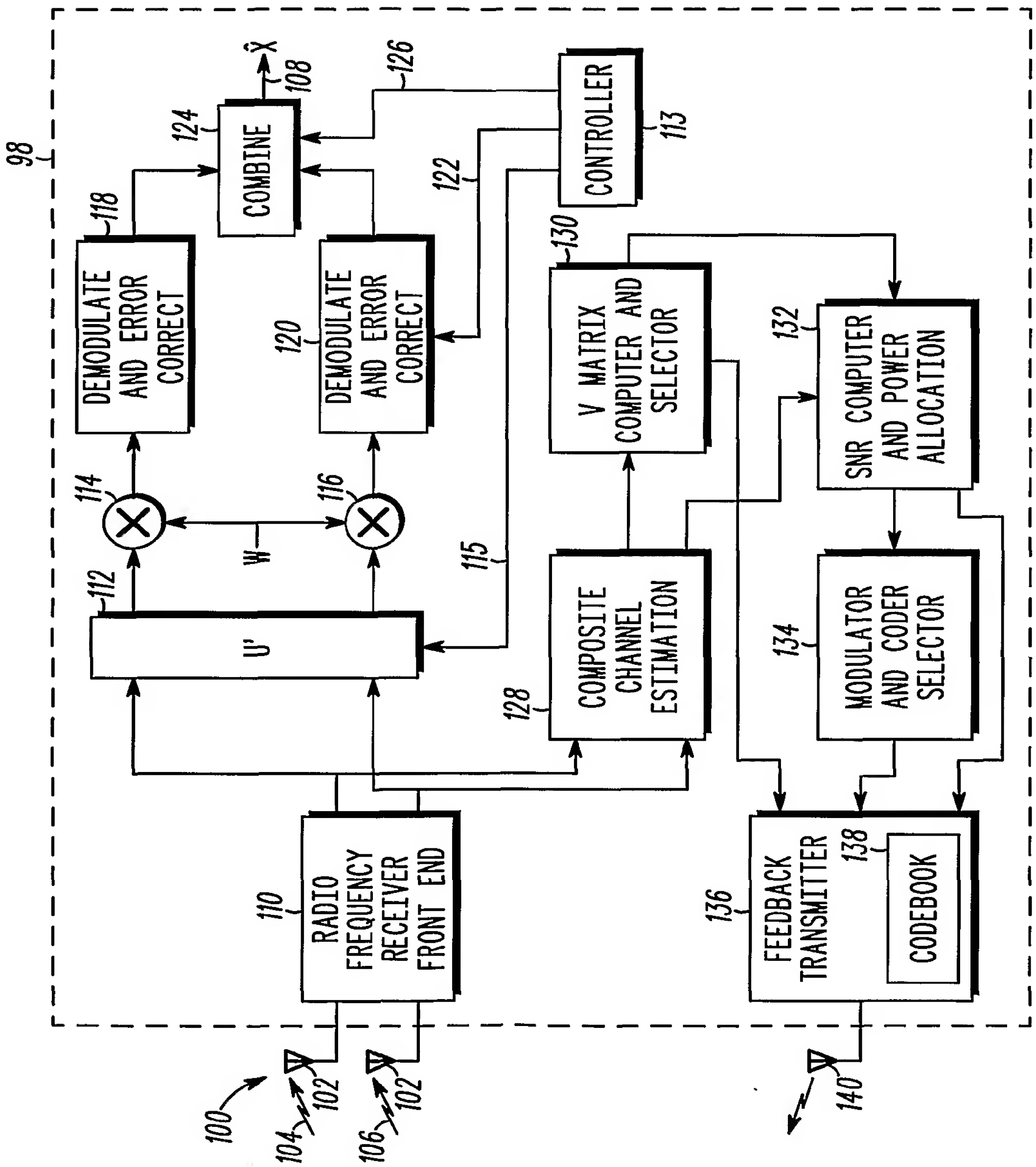



FIG. 4



98
DEMODULATE AND ERROR CORRECT
118
COMBINE
124
$\hat{X}$
108
126
CONTROLLER
113
122
DEMODULATE AND ERROR CORRECT
120
114
116
W
112
U'
115
V MATRIX COMPUTER AND SELECTOR
130
SNR COMPUTER AND POWER ALLOCATION
132
COMPOSITE CHANNEL ESTIMATION
128
MODULATOR AND CODER SELECTOR
134
RADIO FREQUENCY RECEIVER FRONT END
110
FEEDBACK TRANSMITTER
136
CODEBOOK
138
100
102
102
104
106
140


FIG. 5

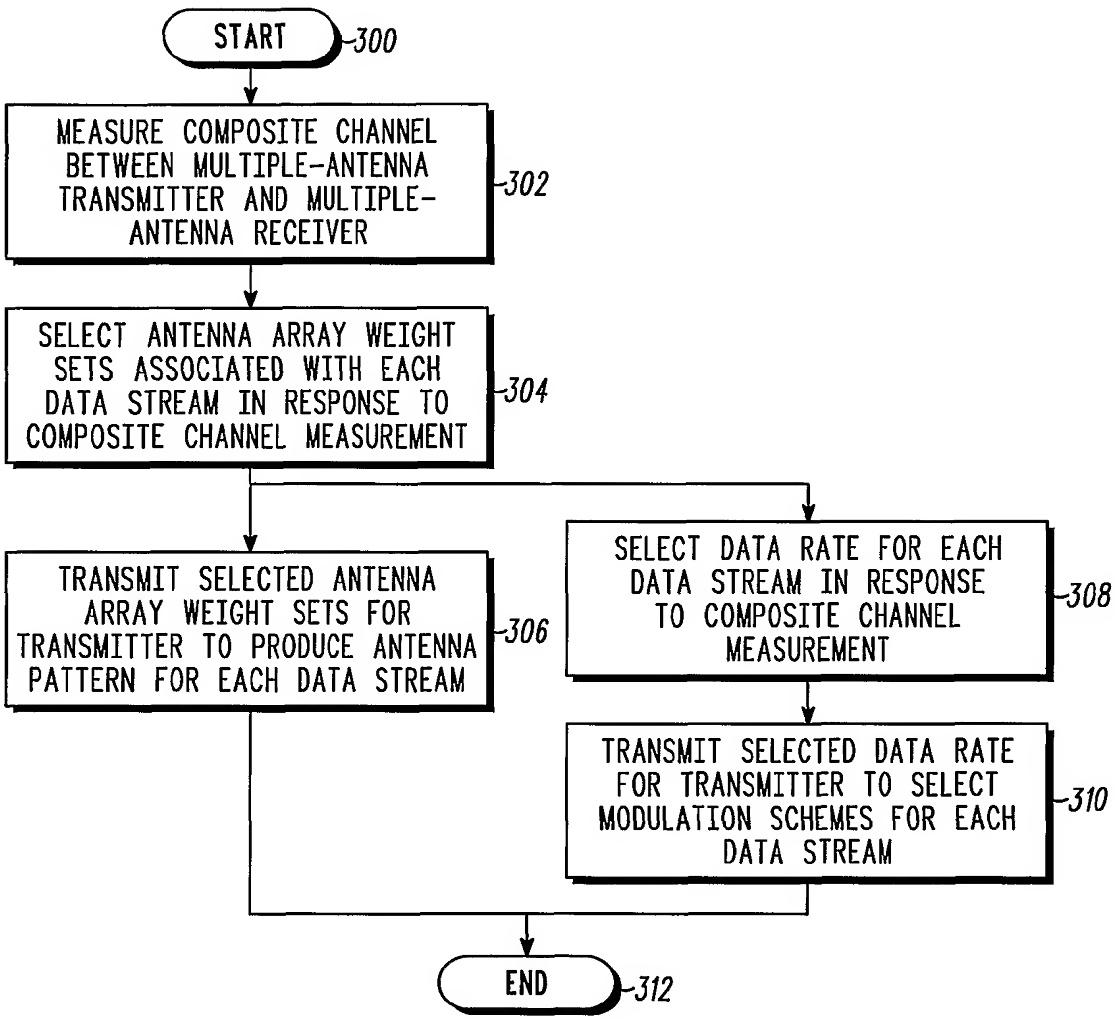

START
300
MEASURE COMPOSITE CHANNEL BETWEEN MULTIPLE-ANTENNA TRANSMITTER AND MULTIPLE-ANTENNA RECEIVER
302
SELECT ANTENNA ARRAY WEIGHT SETS ASSOCIATED WITH EACH DATA STREAM IN RESPONSE TO COMPOSITE CHANNEL MEASUREMENT
304
TRANSMIT SELECTED ANTENNA ARRAY WEIGHT SETS FOR TRANSMITTER TO PRODUCE ANTENNA PATTERN FOR EACH DATA STREAM
306
SELECT DATA RATE FOR EACH DATA STREAM IN RESPONSE TO COMPOSITE CHANNEL MEASUREMENT
308
TRANSMIT SELECTED DATA RATE FOR TRANSMITTER TO SELECT MODULATION SCHEMES FOR EACH DATA STREAM
310
END
312

FIG. 6

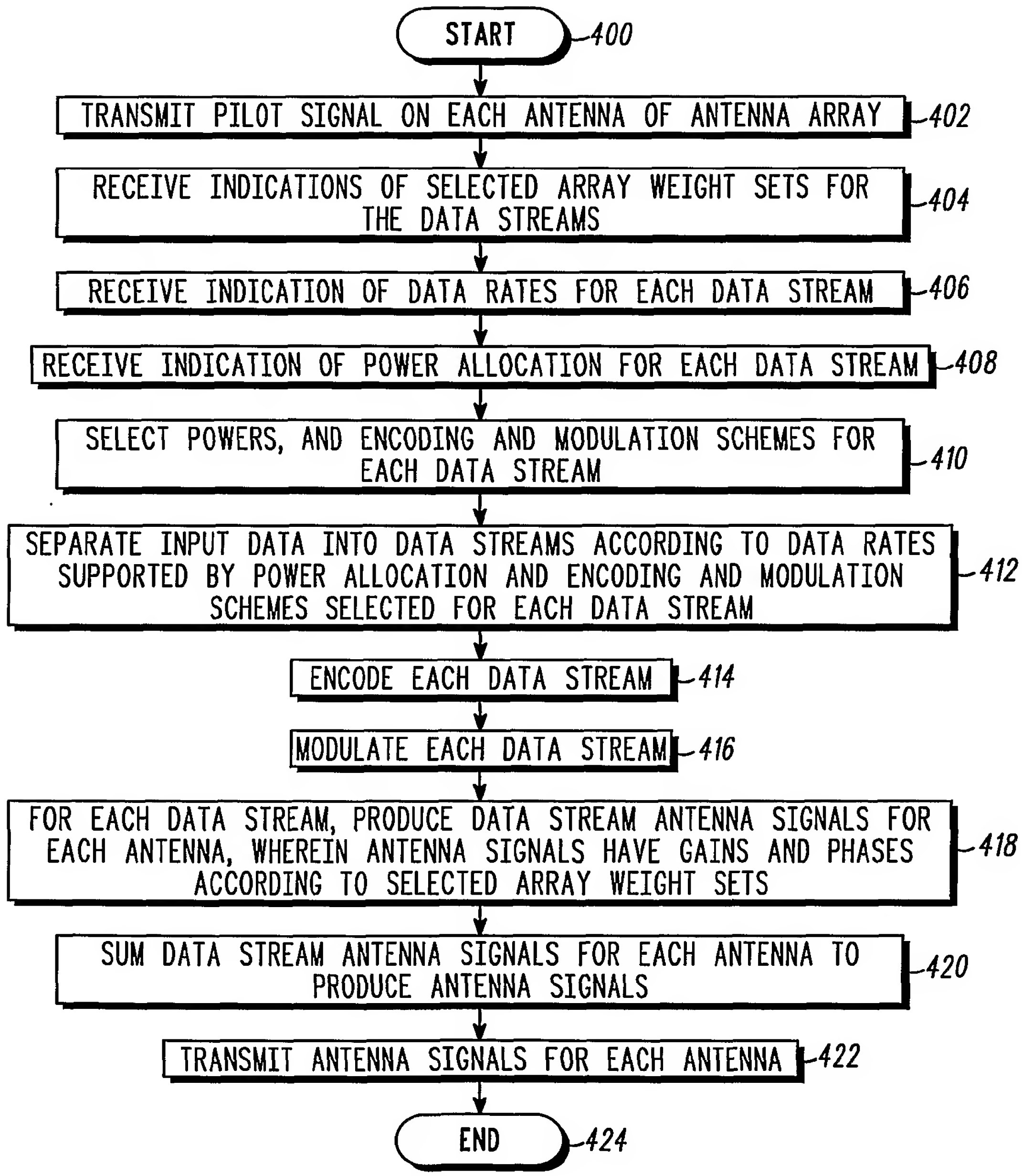

START
400
TRANSMIT PILOT SIGNAL ON EACH ANTENNA OF ANTENNA ARRAY
402
RECEIVE INDICATIONS OF SELECTED ARRAY WEIGHT SETS FOR THE DATA STREAMS
404
RECEIVE INDICATION OF DATA RATES FOR EACH DATA STREAM
406
RECEIVE INDICATION OF POWER ALLOCATION FOR EACH DATA STREAM
408
SELECT POWERS, AND ENCODING AND MODULATION SCHEMES FOR EACH DATA STREAM
410
SEPARATE INPUT DATA INTO DATA STREAMS ACCORDING TO DATA RATES SUPPORTED BY POWER ALLOCATION AND ENCODING AND MODULATION SCHEMES SELECTED FOR EACH DATA STREAM
412
ENCODE EACH DATA STREAM
414
MODULATE EACH DATA STREAM
416
FOR EACH DATA STREAM, PRODUCE DATA STREAM ANTENNA SIGNALS FOR EACH ANTENNA, WHEREIN ANTENNA SIGNALS HAVE GAINS AND PHASES ACCORDING TO SELECTED ARRAY WEIGHT SETS
418
SUM DATA STREAM ANTENNA SIGNALS FOR EACH ANTENNA TO PRODUCE ANTENNA SIGNALS
420
TRANSMIT ANTENNA SIGNALS FOR EACH ANTENNA
422
END
424

*FIG. 7*

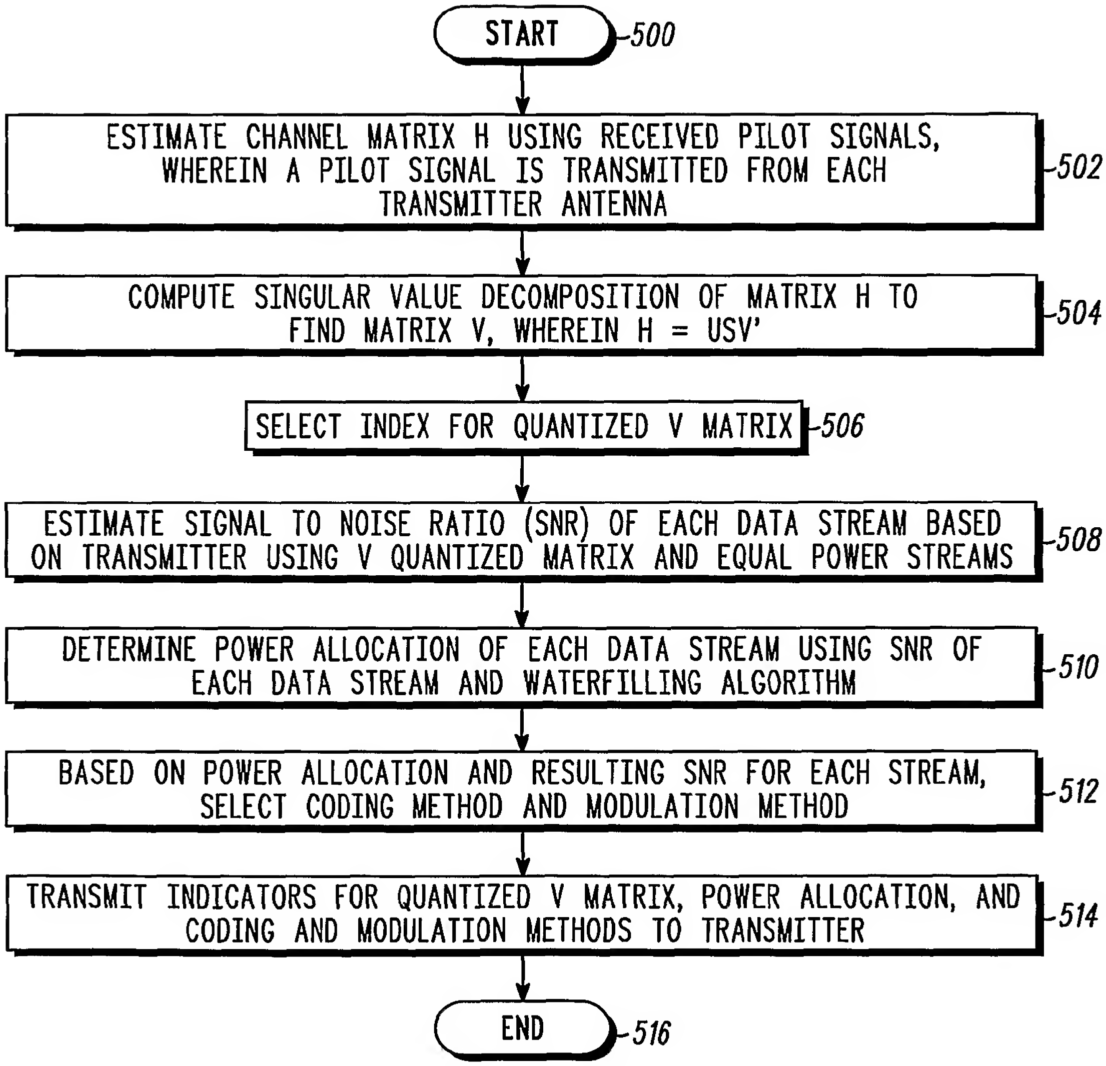

START
500
ESTIMATE CHANNEL MATRIX H USING RECEIVED PILOT SIGNALS, WHEREIN A PILOT SIGNAL IS TRANSMITTED FROM EACH TRANSMITTER ANTENNA
502
COMPUTE SINGULAR VALUE DECOMPOSITION OF MATRIX H TO FIND MATRIX V, WHEREIN H = USV'
504
SELECT INDEX FOR QUANTIZED V MATRIX
506
ESTIMATE SIGNAL TO NOISE RATIO (SNR) OF EACH DATA STREAM BASED ON TRANSMITTER USING V QUANTIZED MATRIX AND EQUAL POWER STREAMS
508
DETERMINE POWER ALLOCATION OF EACH DATA STREAM USING SNR OF EACH DATA STREAM AND WATERFILLING ALGORITHM
510
BASED ON POWER ALLOCATION AND RESULTING SNR FOR EACH STREAM, SELECT CODING METHOD AND MODULATION METHOD
512
TRANSMIT INDICATORS FOR QUANTIZED V MATRIX, POWER ALLOCATION, AND CODING AND MODULATION METHODS TO TRANSMITTER
514
END
516

FIG. 8

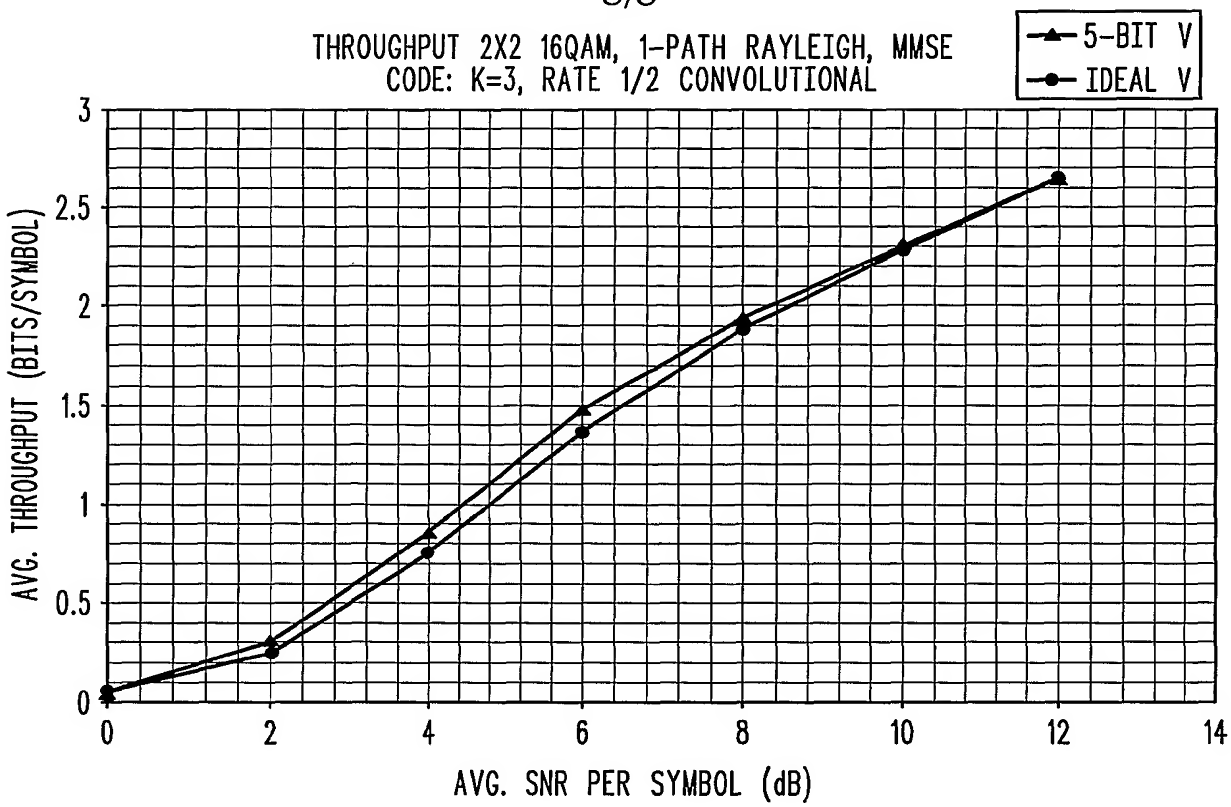
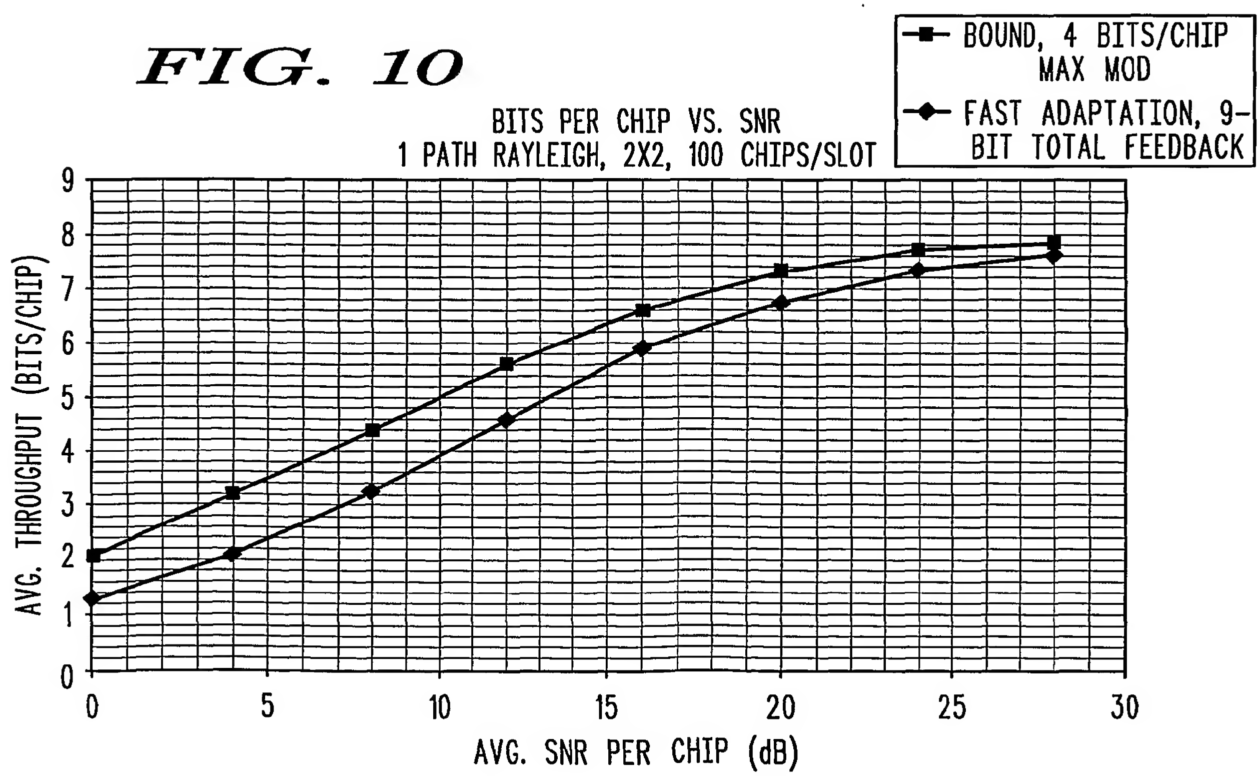

THROUGHPUT 2X2 16QAM, 1-PATH RAYLEIGH, MMSE
CODE: K=3, RATE 1/2 CONVOLUTIONAL
5-BIT V
IDEAL V
AVG. THROUGHPUT (BITS/SYMBOL)
0
0.5
1
1.5
2
2.5
3
0
2
4
6
8
10
12
14
AVG. SNR PER SYMBOL (dB)

*FIG. 9*

*FIG. 10*


BITS PER CHIP VS. SNR
1 PATH RAYLEIGH, 2X2, 100 CHIPS/SLOT
BOUND, 4 BITS/CHIP MAX MOD
FAST ADAPTATION, 9-BIT TOTAL FEEDBACK
AVG. THROUGHPUT (BITS/CHIP)
0
1
2
3
4
5
6
7
8
9
0
5
10
15
20
25
30
AVG. SNR PER CHIP (dB)

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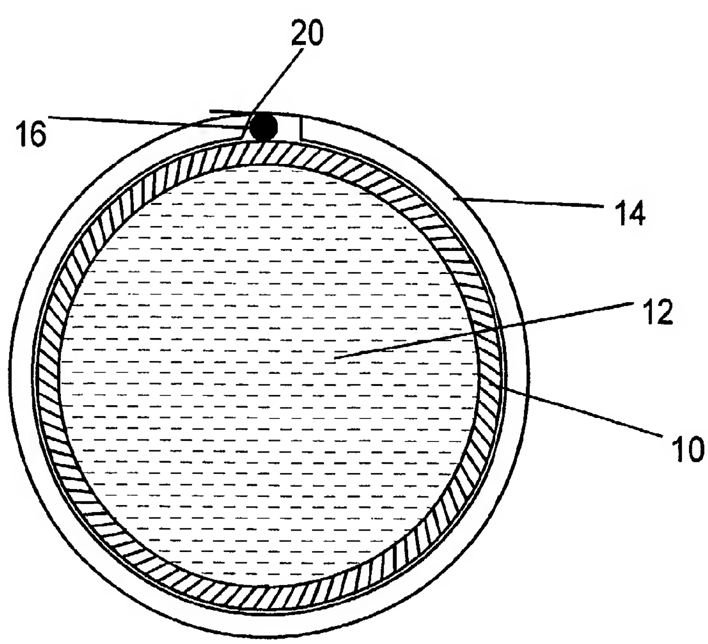


**(57) Abstract:** Leaks are detected by wrapping a vessel such as a pipe (10) or tank (28) in a skin (14, 14A, 30) which traps escaping fluid (12) at least long enough to direct it, in the vicinity of the leak, towards a sensor line (16) employing fibre optics to detect the fact of a leak and to detect how far along the a fibre optic line the leak exists. The skin (14, 14A, 30) can be longitudinally applied or wrapped onto the vessel (10, 28). Bindings (24) can be used to attach the skin (14, 14A, 30) to the outside of the vessel (10, 28). Sensor line (16) couplings (26) can be employed between lengths of pipe (10) to create monitored sections of pipe (10) which can be joined together. Sensor lines (16) can be applied to the outer surface of a vessel (10, 28) and covered with the skin (14, 14A, 30). Sensor lines (16) can be stuck or woven into the skin (14, 14A, 30) incorporates elastic ridges (18, 18A, 18C) which face the vessel and direct escaping fluid towards the sensor line (16). A control system (46, 40, 48, 44) is provided to shut down a tank (28) or pipeline (36, 38), at least in the vicinity of a leak, if a leak is detected, and can include shutting down pumps (40), closing valves, and voiding items (10, 28) subject to the leak.

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# Method and Apparatus for Leak Detection and Location

The present invention relates to a method and apparatus for detecting leaks. The invention particularly relates to detecting fluid (liquid or gaseous) leaks in vessels such as pipelines and storage tanks.

Pipelines are vessels used for conducting fluids, such as gas, water, chemicals or oil, from geographical place to geographical place, or, in an industrial setting, between tanks, between processes, or between processes and tanks and vice versa. Tanks are vessels used for temporary or permanent bulk storage, where the fluid enjoys at least a temporary period of non-movement. There is often a need to detect whether or not a pipe or tank is leaking. Often, this is done by visual inspection, by which time any damage done by the escaping fluid has already occurred, or is done by flow measurements where it is noted that the ingress of fluid volume or quantity is greater than the egress of fluid volume or quantity. In a tank, a leak becomes apparent either by inspection or by noting that the content of the tank has decreased over the sum of its inflow and outflow. The prior art is silent upon any method which will detect a leak, as it occurs anywhere in the system or tank under surveillance. The prior art is also silent upon any method automatically to locate the position of a leak as it occurs.

The present invention seeks to provide a method and apparatus for the rapid detection of the occurrence of a leak anywhere in a vessel or system of vessels under surveillance and the rapid determination of the location of the leak. The present invention also seeks to provide a method and apparatus whereby a leak, anywhere in a system of vessels under surveillance, can be shut down at the moment of detection.

According to a first aspect, the present invention consists in an apparatus for determining the occasion and location of a fluid leak in a vessel, said apparatus comprising: a sensor line, on the outside of the vessel, for detecting where, along the length of the sensor line the sensor line is in contact with the fluid; and a skin, for use on the outside of the vessel and operative to direct leaked fluid towards the sensor line in the vicinity of any leak.

According to a second aspect, the present invention consists in a method for determining the occasion and location of a fluid leak in a vessel, said method including the steps of: disposing a sensor line, on the outside of the vessel, for detecting where, along the length of the sensor line the sensor line is in contact with the fluid; disposing a skin on the outside of the vessel; and employing said skin to direct leaked fluid towards the sensor line in the vicinity of any leak.

The invention further provides that the said skin is operative, at least temporarily, to contain fluid, leaked from the vessel.

The invention further provides that the skin is operative to enclose both the vessel and the sensor line.

The invention further provides that the vessel can be a pipe, that the sensor line is disposable longitudinally along the pipe, that the skin comprises elastic ridges, to be pressed against the outer surface of the pipe, that the elastic ridges are operative to inhibit fluid flow in a longitudinal direction along the outside of the pipe, and that the elastic ridges are operative to direct fluid flow on the outside of the pipe in a circumferential direction towards the sensor line.

The invention further provides that the skin can be disposed longitudinally along the outside of the pipe and can be closed, longitudinally, by means of a securing cover

The invention further provides that the skin can be helically wrapped about the pipe.

The invention further provides that the skin can be further fixed onto the outside of the pipe by means of spaced bands.

The invention further provides that the vessel can be a tank, that the sensor line can be disposed on the outer surface of the tank, that the skin can comprise elastic ridges, to be pressed against the outer surface of the tank, that the elastic ridges can be operative to form containment zones to contain and accumulate fluid from any leak for the fluid to come into contact with the sensor line.

The invention further provides that the skin can be wrapped around the outside of the tank in a close helix, that the skin can comprise partial containment zones, and that a partial containment zone in one coil of the helix can co-operate with a partial containment zone in an adjacent coil of the helix to contain and accumulate fluid from any leak for the fluid to come into contact with the sensor line.

The invention further provides that the sensor line can be incorporated into the fabric of the face of the skin which is for presentation to the outside surface of the vessel.

The invention further provides that the sensor line can be a fibre optic line.

The invention is further explained, by way of example, by the following description, in conjunction with the appended

drawings, in which:

Figure 1 shows a cross sectional view of a vessel in the form of a pipe to which the present invention has been applied.

Figure 2 shows an opened out view of the skin enclosing the pipe of figure 1.

Figure 3 shows one way of attaching the skin, of Figures 1 & 2, to a pipe.

Figure 4 shows another way that a skin may be applied to a pipe.

Figure 5 is an example of the skin, which can be used in Figure 4.

Figure 6 is an angled view of the pipe, shown in Figure 1.

Figure 7 illustrates how sensor lines may be incorporated into the skin otherwise shown in Figure 2.

Figure 8 is an example of how sensor lines may be incorporated into the skin, otherwise shown in Figure 5.

Figure 9 is a drawing of a vessel in the form of a tank showing how a skin may be applied thereto, not only to detect a leak, but to determine at what part of the tank the leak is occurring.

Figure 10 is a view of the ridge structure of a skin suitable for use on the tank of Figure 9, and showing how sensor lines may be applied thereto.

Figure 11 is a cross sectional view of a pipeline showing how sensor lines can be attached other than at the top of a pipe and how the skin need not provide containment of leaking fluid, merely direction toward the sensor line.

And

Figure 12 is a projected schematic view of an exemplary pumping system, according to the present invention, showing a control system suitable for monitoring a pipeline for leaks, for shutting down the pipeline when a leak is detected, and for providing rapid indication of the location of the leak.

Attention is drawn to Figure 1. A pipe 10, carrying a fluid load 12, is surrounded by a containment skin 14. Within the containment skin 14, and on top of the pipe 10, a sensor line 16 is provided.

The fluid load 12 can be gaseous or liquid. It can consist of chemical gases, fuel gases, hydrocarbons, oil, water, food stuffs such as milk, chemical liquids and, indeed, just about any type of thing that can be driven along a pipe 10.

Part of the purpose of the skin 14 is to protect the sensor line 16 when the pipe 10 is buried in the ground, encased in concrete, or otherwise exposed to a harsh surrounding environment. Another purpose of the containment skin 14 is, at least temporarily, to contain any of the fluid load 12 which may escape from the pipe 10 at least long enough to duct the escaping fluid load 12 towards the sensor line 16.

The sensor line 16 is, in the preferred embodiment of this invention, a fibre optic line which, as is well known to those skilled in the art, can be adapted and used to detect, among other things, moisture, specific chemicals, changes in

temperature, oil and natural gas. The present invention can also employ any other elongated sensor or array of sensors, including spaced gas, chemical oil, temperature and other sensors. More than one sensor line 16 can be provided.

By placing the sensor line 16 on top of the pipe 10, the sensor line 16 avoids contact with accumulated contaminants and debris which might accrete in the bottom of the skin 14, and ensures, thereby, that the sensor line 16 responds only to true leaks. As will become clear from the description of Figure 11, other arrangements are possible within the scope of the present invention.

Attention is drawn to Figure 2, showing the face of the skin 14 applied to the pipe 10. The reverse face of the skin 14 is smooth. The skin 14 comprises elastic ridges 18 which are wrapped circumferentially around the pipe 10 and held in place by a securing cover 20 which can be used to close the skin 14 by means of adhesives or other gripping means, and also serves to protect the sensor line 16 and to maintain it in position on top of the pipe 10.

Should a leak occur, the circumferential elastic ridges 18, against the face of the pipe 10, prevent the escaping fluid 12 from moving longitudinally along the pipe 10. Any fluid escape is ducted substantially circumferentially around the pipe at the location where it occurred. As is known in the art, a fibre optic sensor line 16 can detect the position along its length where interaction with a selected or detectable fluid has occurred. By confining the leaking fluid 12 to the point where the leak occurred, at least long enough for the escaping fluid to encounter the sensor line 16, and by sensing the position of interaction of the fibre optic sensor line 16 with the fluid, ducted towards the fibre optic sensor line 16, it is possible to obtain a very rapid detection of

the fact that a leak has occurred and to find the position of that leak, with great accuracy. Distance along the fibre optic sensor line 16 is measured by finding the time delay for light travelling along the fibre optic sensor line 16.

The lower portion of Figure 2 is a side view of the skin 14, looking in the direction of the arrow 22, showing the elastic ridges 18 in profile.

Attention is drawn to Figure 3 showing one way in which the skin 14 of Figure 1 can be attached to a pipe 10 (shown in phantom outline) by means of spaced bands 24 braced over the skin 14 at intervals along the pipe 10, not only to hold the skin 14 onto the pipe 10 but also to improve the ability of the skin 10 longitudinally to trap escaping fluid from the pipe 10. Couplings 26 allow the sensor line 16 to be coupled to sensor lines 16 on adjacent pipes 10. The arrangement shown in Figure 3 constitutes a complete and portable pipe 10 unit which can be moved and installed as an entirety.

Attention is drawn to Figure 4 showing another way in which a spiral skin 14A can be helically wrapped around the pipe 10 (shown in phantom outline).
Figure 5 shows the arrangement of elastic ridges 18A on the spiral skin 14A. It is perceived that the elastic ridges 18A are longitudinal in the sense of direction of the spiral skin 14A, and when wrapped around the pipe 10 in a close fitting helical fashion, the elastic ridges 18A in the spiral skin 14A form a more or less vertical (circumferential) pattern which contains any leak in the vicinity of the sensor line 16. In the example shown in Figure 4, the sensor line 16 is simply placed on top of the pipe 10 and the spiral skin 14A wrapped around the pipe 10. As will later be seen, better arrangements than this can be made. The elastic ridges 18, 18A can be provided at various angles to the longitudinal

direction of the skin 14, dependently upon the intended manner of attachment thereof to the pipe 10.

The arrangement shown in Figure 4 can further be improved by, in addition to helical wrapping, employing spaced bands 24 as illustrated in Figure 3.

Attention is drawn to Figure 6, showing, for clarity, a projected view of the cross section of Figure 1, and showing, in particular, how the sensor line 16 is enclosed by and protected by the securing cover 20. The containment skin 14, shown in Figure 6, is shown cut away to cover only a portion of the surface of the pipe 10 so that the disposition of the sensor line 16 can be seen. While the sensor line 16 is shown covered by the securing cover 20, it is to be understood that the skin 14 can be secured to the pipe 10 by other means and the sensor line 16 left uncovered by the skin 14 simply to have any escaping fluid 12 ducted in its direction.

Attention is drawn to Figure 7 showing one way in which one or more sensor lines 16 can be threaded, permanently, through the material or fabric on the inside of the containment skin 14, otherwise shown in Figure 2. If the inner face of the containment skin 14 contains any woven fabric element, the sensor line 16 can simply be woven into the fabric. Otherwise, the sensor line 16 can simply be moulded into or attached to the inner surface of the containment skin 14.

Figure 8 shows a manner in which sensor lines 16 can be applied as an integral part of the spiral skin 14A shown in Figures 4 and 5. One or more sensor lines 16 are placed on the inner surface of the spiral skin 14A to run in the spaces between the elastic ridges 18A in the spiral skin 14A. They can be incorporated in just the same manner as was earlier described for Figure 7.

Attention is next drawn to Figure 9 showing an application of the present invention to a tank 28 containing a fluid. A tank skin 30 is wrapped around the tank 28, here shown wrapped in a spiral manner but other manners are possible, to cover the surface of the tank 28 as much as possible. The sensor line 16 leads all around the helical wrapping of the tank skin 30 and is accessible at either end via couplings 26A.

Attention is drawn to Figure 10 showing the face of the tank skin 30 which faces the tank 28. Elastic ridges 18C form entire containment zones 32 in the centre of the tank skin 30 and partial containment zones 34 at the sides thereof and extending on either side, away from the entire containment zone 32 to the edge of the tank skin 30. When pressed against the tank 28, the entire containment zones 32 keep any escaping fluid, at least temporarily, from moving. The partial containment zones 34 co-operate with partial containment zones in adjacent wraps of the tank skin 30 to form at least a temporary containment area for escaping fluid. The sensor line 16, in the example shown, is provided only through the entire containment zone 32 and through one of the partial containment zones 34. As the tank skin 30 is wrapped, the sensor line 16 in the adjacent wrap of tank skin 30 acts to provide a sensor line 16 in a co-operation of partial containment zones where one is not present on the adjacent tank wrap 30. Since it is possible to measure the distance along the fibre optic sensor line 16, it is possible to measure where, on the surface of the tank 28 a leak has occurred. By knowing which portion of the tank 28 is covered by which portion of the tank skin 30, the location of a leak can be rapidly determined.

Attention is next drawn to Figure 11, showing the cross sectional pipe arrangement of Figure 1, but with a different arrangement for the relative position of the sensor line 16.

The sensor line 16 is, for preference, provided at the top of the pipe 10. However, as is clear if more than one sensor line is used, the sensor line 16 may be otherwise disposed on the pipe 10. Figure 11 shows the sensor line 16 in a first position near the base of the pipe 10. The sensor line 16 is also shown in a second position where it is part way up the pipe 10. The sensor line can be attached, in varying positions on the pipe 10 by means of adhesives, tapes and spaced bands. All that is important is that the sensor line is in a position to interact with the escaping fluid 12, should an escape occur. The sensor line 16 can be longitudinally disposed along the pipe, or can be spirally wound around the pipe 10 in an arrangement incorporating the arrangement of the spiral skin 14A of Figure 4 with an incorporated sensor lines 16 or lines 16 as shown in Figure 7 and Figure 8.

The elastic ridges 18 18A, shown in Figures 2, 5, 7, 8 and 10, can also be otherwise provided, according the present invention. The ridges 18 18A 18C may restrict the escaping fluid 12 from longitudinal migration along the pipe 10 . However, localisation of the escaping fluid 12 is only necessary for enough time for the fluid 12 to reach the sensor line16 . Thus, arrangements of ridges 18 18A 18C, right down to there being no ridges 18, 18A 18B present, merely a conformal skin that spreads the escaping fluid across the surface of the pipe 10 at least as far as the sensor line, can also be provided and can work according the the present invention.

The skin 14 need not contain escaping fluid. The invention also functions if the fluid 12 can escape. The skin, in Figure 11, shows a means of egress for the fluid, just to emphasize this point, where the securing cover 20 does not provide a fluid tight enclosure around the pipe 10.

Attention is drawn to Figure 12 showing a projected schematic diagram illustrating how the present invention provides a control feature for tanks or pipelines.

An exemplary pipeline 36 comprises one or more pipeline sections 38 between pumping stations 40 through which a fluid is propelled as illustrated by arrows 42. The pipeline 36 could equally well be a tank.

A sensor line 16D is provided for leak detection on one or more adjacent pipeline or tank sections 38 and is driven and monitored by a sensor driver 44 which provides laser light, laser detectors, timers and all the other apparatus which, as is already known in the art, is necessary for the detection and location of a fluid leak. A monitor 46 receives output from the sensor driver 44 and displays the current state of the monitored pipeline 36 or tank . As soon as the monitor 46 detects that a leak has occurred, it sends an operating signal to a pump controller 48 which sends a control signal to each pumping station 40 on the monitored pipeline 38 or tank causing each pumping station 40 to shut down. The monitor 46 can then provide humanly interpretable input for assessing the progress of leak repair and recovery.

The pump controller 48 could equally receive its operating signal directly from the sensor driver. The operating signal for the pumping stations 40 can be provided to all pumping stations 40 on the monitored and controlled pipeline 36or tank, or can be provided only to that pumping station 40 or those pumping stations 40 which is or are nearest to and contain the loss of fluid from the leak. In this example, the pipeline 36 or tank is provided with pumping stations. The invention provides that a pipeline or tank can comprise pumping stations, stop valves, and, indeed, any device which can be applied or ceased to be used in order to shut down the

loss of fluid flow from the pipeline 36 or tank whenever a leak is detected. This can comprise shutting down all flow. It can also comprises starting fluid movement out of the damaged section to prevent leakage loss, or a combination of both techniques. Thus, one pumping station 40 can be shut down in a pipeline 36 upstream of a leak and flow towards the leak stopped, while pumping downstream of the leak can be continued or enhanced to empty the pipeline 36. A monitored tank can have its inflow stopped while its outflow continues until the tank is empty, or until leakage stops as, for example, when the level in the lank falls below the height of the leak of until leakage is no longer detected.

The invention also provides that the sensor driver 44 can drive and monitor more than one sensor line 16D, either in the same pipeline (or tank) section 38 (where more than one indication of a leak can be employed to confirm a leak and prevent falsely indicated shutdowns) or in different pipeline (or tank) sections 38.

While it is implicit in the disclosure of the invention, it is here stated, for clarity, that the skin or skins 14 14A 30 can be flexible for wrapping around pipes, tanks and any other vessels to which the invention can be applied and that the invention, as described and claimed, can be retrofitted to existing pipes, tanks and other vessels.

The invention has so far been described by way of example. The invention is further described by the following claims.

Claims

1. An apparatus for determining the occasion and location of a fluid leak in a vessel, said apparatus comprising: a sensor line, on the outside of the vessel, for detecting where, along the length of the sensor line the sensor line is in contact with the fluid; and a skin, for use on the outside of the vessel and operative to direct leaked fluid towards the sensor line in the vicinity of any leak.

2. An apparatus, according to claim, wherein said skin is operative, at least temporarily, to contain fluid, leaked from the vessel.

3. An apparatus, according to claim 2, wherein said skin allows the escape of said fluid.

4. An apparatus, according to claims 1, 2 or 3, wherein said skin is operative to enclose both said vessel and said sensor line.

5. An apparatus, according to claims 1, 2, 3 or 4 for use where said vessel is a pipe, wherein said sensor line is disposable along the pipe

6. An apparatus, according to claim 5, wherein said skin comprises elastic ridges, to be pressed against the outer surface of the pipe and wherein said elastic ridges are operative to inhibit fluid flow in a longitudinal direction along the outside of the pipe and to directing fluid flow on the outside of the pipe towards said sensor line.

7. An apparatus, according to claim 5 or 6, wherein said skin is disposable longitudinally along the outside of the

pipe and closable, longitudinally, by means of a securing cover

8. An apparatus, according to claim 5 or 6, wherein said skin is helically wrappable about the pipe.

9. An apparatus according to claims 5, 6, 7 or 8, wherein said skin is further fixable onto the outside of the pipe by means of spaced bands.

10. An apparatus according to claims 1, 2, 3 or 4 for use where said vessel is a tank, wherein said sensor line is disposable on the outer surface of the tank.

11. An apparatus, according to claim 10, wherein said skin comprises elastic ridges, to be pressed against the outer surface of the tank, and wherein said elastic ridges are operative to form containment zones to at least temporarily contain fluid from any leak for the fluid to come into contact with said sensor line.

12. An apparatus, according to claims 10 or 11, wherein said skin is wrappable around the outside of the tank in a closed helix, wherein said skin comprises partial containment zones, and wherein a partial containment zone in one coil of the helix is co-operative with a partial containment zone in an adjacent coil of the helix to at least temporarily contain fluid from any leak for the fluid to come into contact with said sensor line.

13. An apparatus, according to any of the preceding claims, wherein said sensor line is incorporated into the fabric of the face of said skin which is for presentation to the outside surface of the vessel.

14. An apparatus, according to any of the preceding claims, wherein said sensor line is a fibre optic line.

15. An apparatus, according to any one of the preceding claims, including monitoring means, responsive to said sensor line to shut down said vessel in the event of detection of a leak.

16. An apparatus, according to claim 15, wherein said monitoring means is operative to cease pumping fluid at least into that portion of the vessel subject to the leak.

17. An apparatus, according to claims 15 or 16, wherein said monitoring means is operative to shut off at least that portion of the vessel subject to the leak.

18. An apparatus, according to any one of claims 15 to 17, wherein said monitoring means is operative to empty at least that portion of the vessel subject to the leak.

19. A method for determining the occasion and location of a fluid leak in a vessel, said method including the steps of: disposing a sensor line, on the outside of the vessel, for detecting where, along the length of the sensor line the sensor line is in contact with the fluid; disposing a skin on the outside of the vessel; and employing said skin to direct leaked fluid towards the sensor line in the vicinity of any leak.

20. A method, according to claim 19, including the step of employing said skin, at least temporarily, to contain fluid, leaked from the vessel.

21. A method according to claim 19, including the step of allowing the escape of said fluid.

22. A method, according to claims 19, 20 or 21, including the step of employing said skin to enclose both said vessel and said sensor line.

23. A method, according to any one of claims 19 to 22, where said vessel is a pipe, including the step of disposing said sensor line along the pipe.

24. A method, according to claim 23, including the steps of: employing elastic ridges on said skin; and pressing said elastic ridges against the outer surface of the pipe to inhibit fluid flow in a longitudinal direction along the outside of the pipe and to directing fluid flow on the outside of the pipe towards said sensor line.

25. A method, according to claims 23 or 24, including the steps of: disposing said skin longitudinally along the outside of the pipe; and closing said skin, longitudinally, by means of a securing cover

26. A method, according to claims 23 or 24, including the step of helically wrapping said skin about the pipe.

27. A method, according to any one of claims 23, 24, 25 or 26, including the step of fixing said skin onto the outside of the pipe by means of spaced bands.

28. A method, according to claims 19, 20, 21 or 22, where said vessel is a tank, said method including the steps of: disposing said sensor line on the outer surface of the tank.

29. A method, according to claim 28, including the steps of: employing elastic ridges on said skin; and pressing said elastic ridges against the outer surface of the tank to form containment zones to contain and accumulate fluid from any leak for the fluid to come into contact with said sensor line locally to the leak.

30. A method, according to claims 28 or 29, including the steps of: wrapping said skin around the outside of the tank in a close helix; and employing partial containment zones in said skin, a partial containment zone in one coil of the helix being co-operative with a partial containment zone in an adjacent coil of the helix to at least temporarily contain fluid from any leak for the fluid to come into contact with said sensor line locally to the leak.

31. A method, according to any one of claims 19 to 30, including the step of including said sensor line in the fabric of the face of said skin which is for presentation to the outside surface of the vessel.

32. A method, according to any one of claims 19 to 31, including the step of employing, in said sensor line, a fibre optic line.

33. A method, according to any one of claims 19 to 32, including the steps of monitoring said sensor line; and shutting down said vessel in the event of detection of a leak.

34. A method, according to claim 30, wherein said step of shutting down said vessel includes the step of ceasing to pump fluid at least into that portion of the vessel subject to the leak.

35. A method, according to claims 33 or 34, wherein said step of shutting down said vessel includes the step of shutting off at least that portion of the vessel subject to the leak.

36. A method, according to any one of claims 27 to 29, wherein said step of shutting down said vessel includes the step of emptying at least that portion of the vessel subject to the leak.

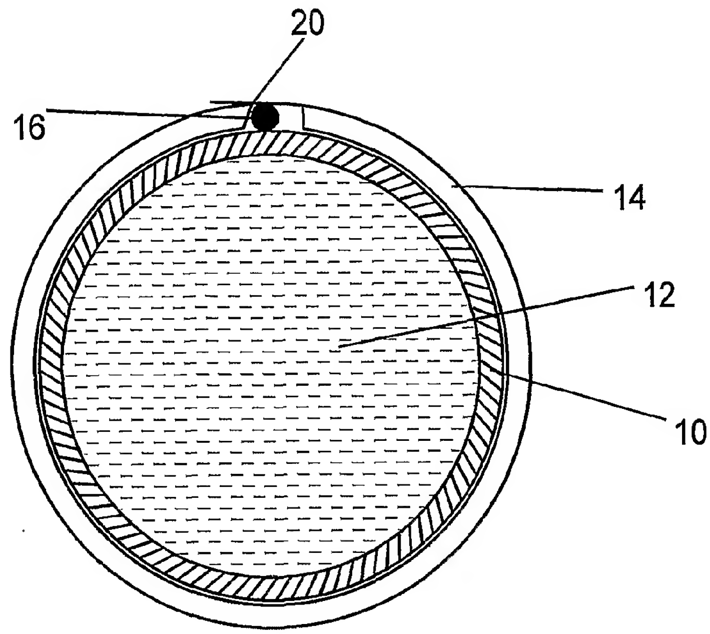


FIGURE 1 α

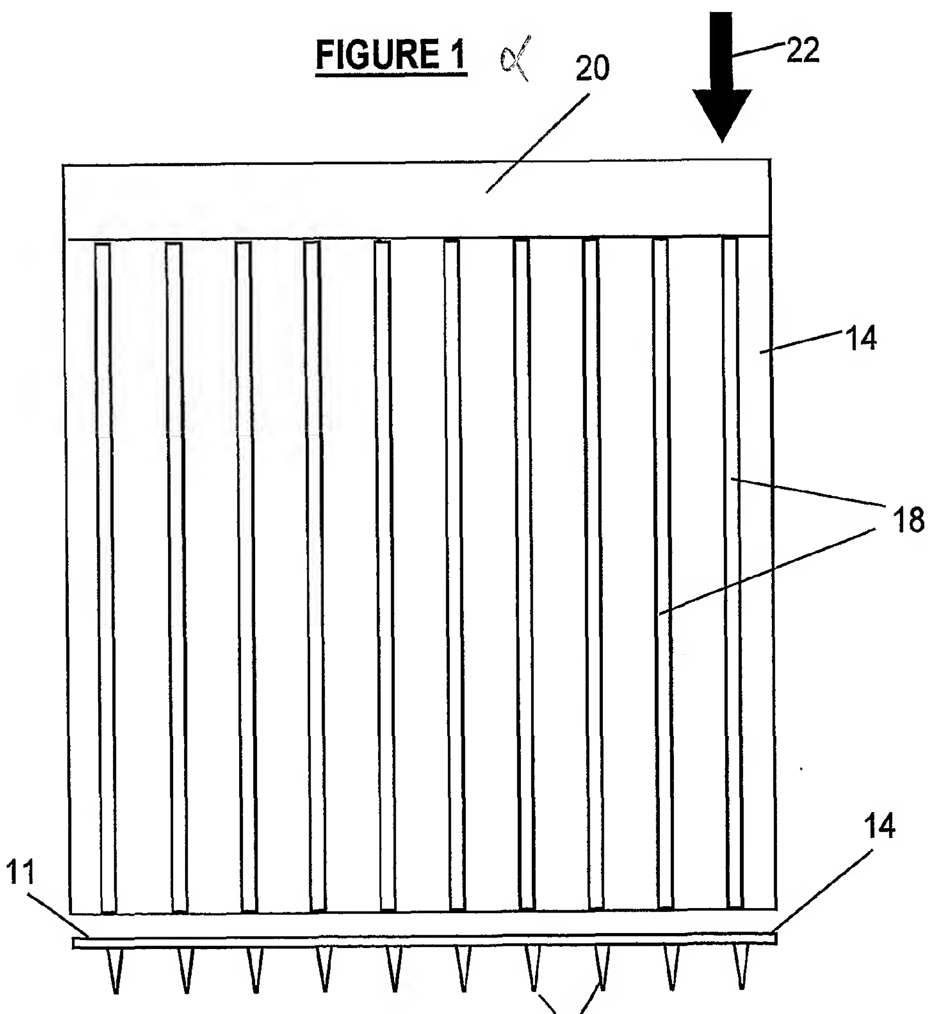


FIGURE 2

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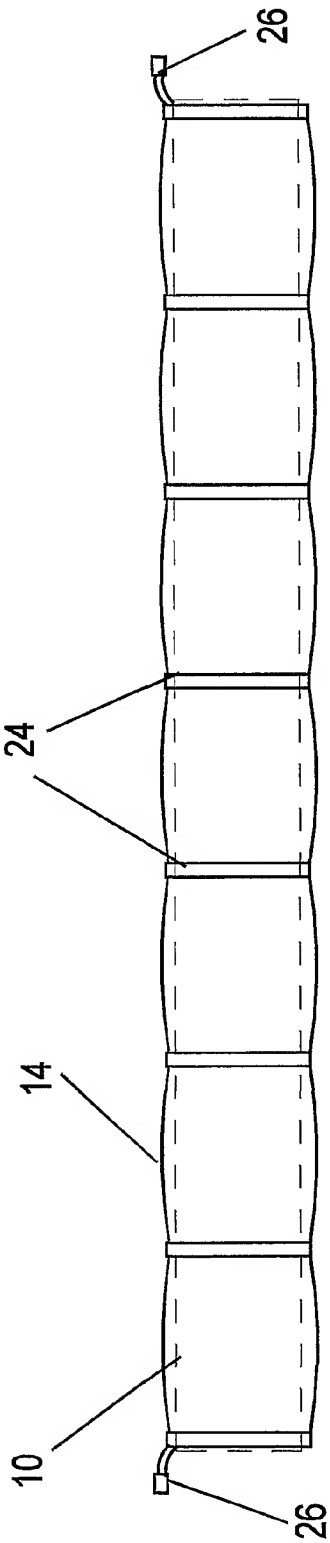


FIGURE 3

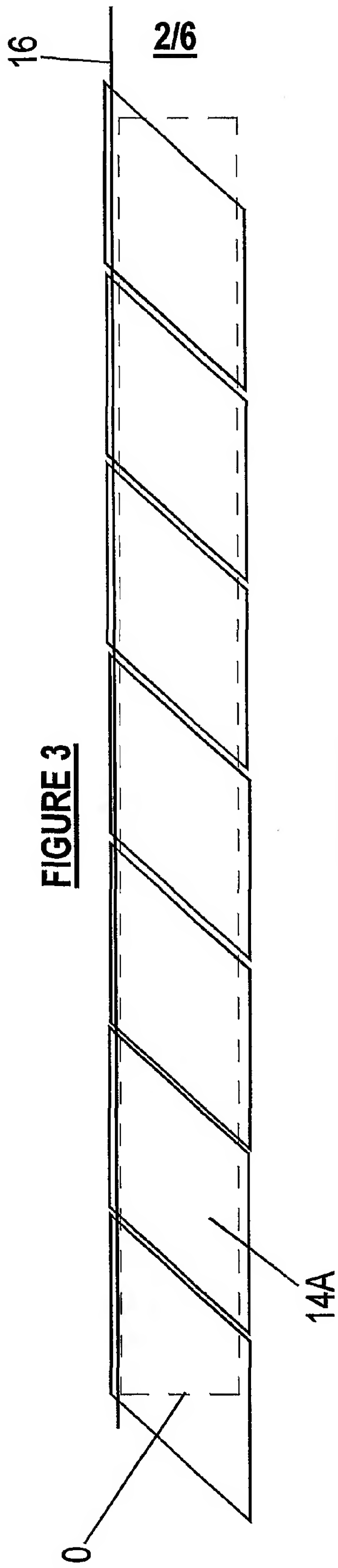


FIGURE 4

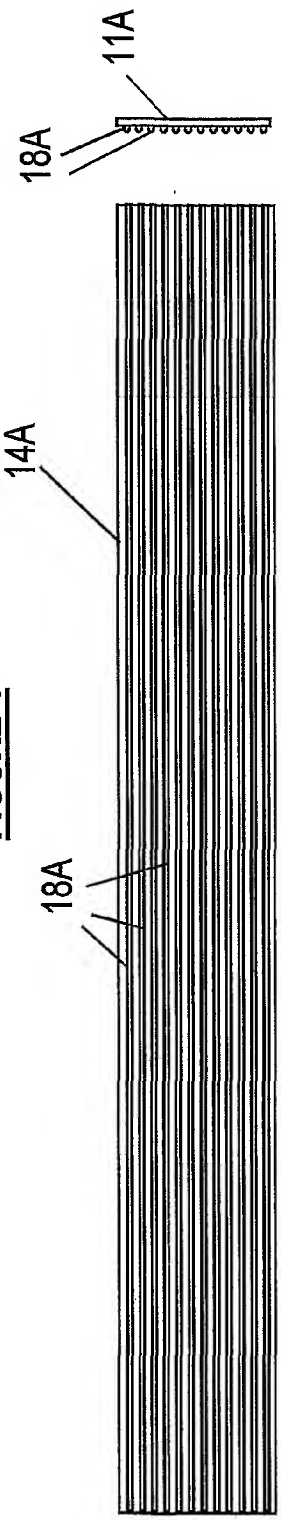


FIGURE 5

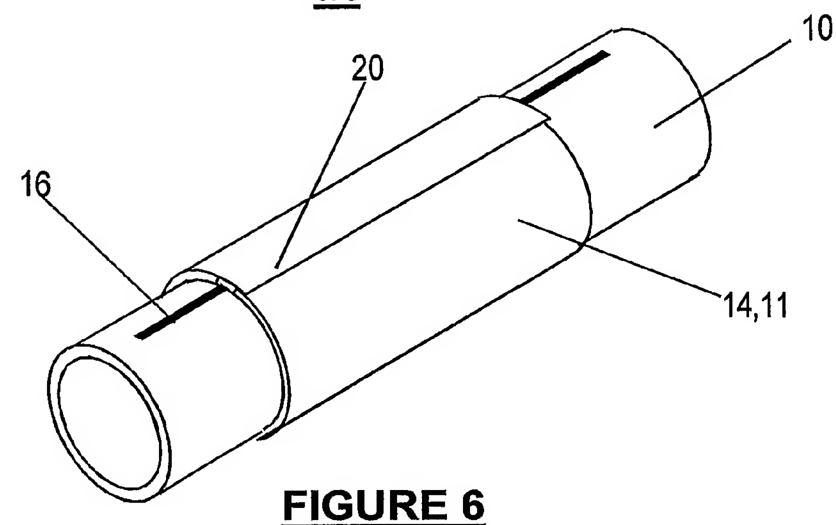


FIGURE 6

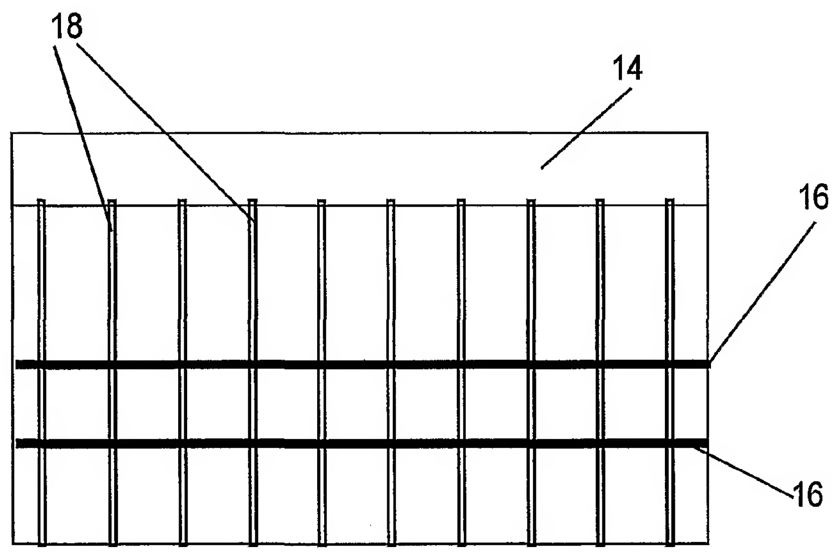


FIGURE 7

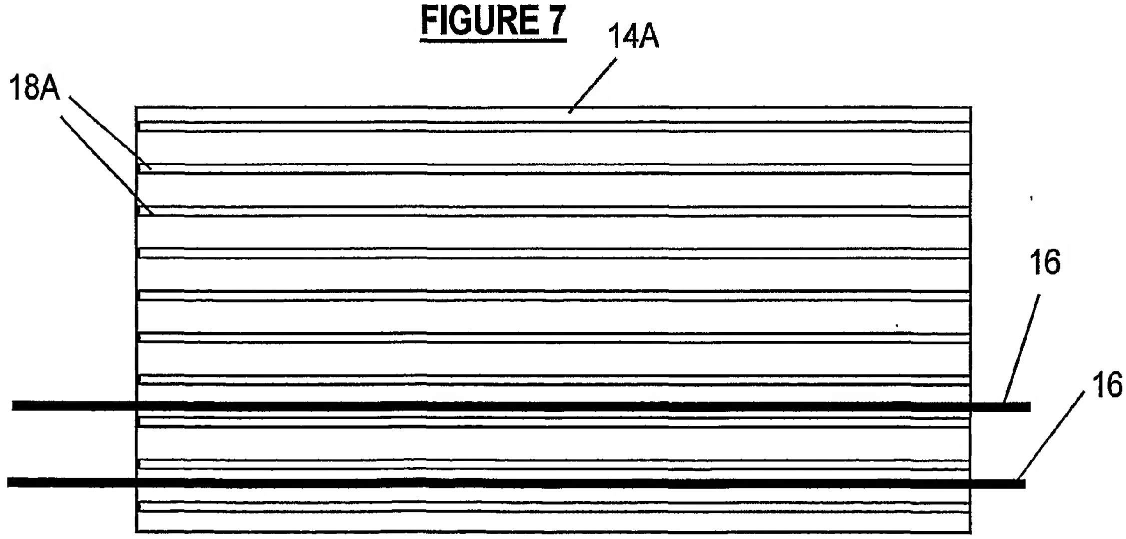


FIGURE 8

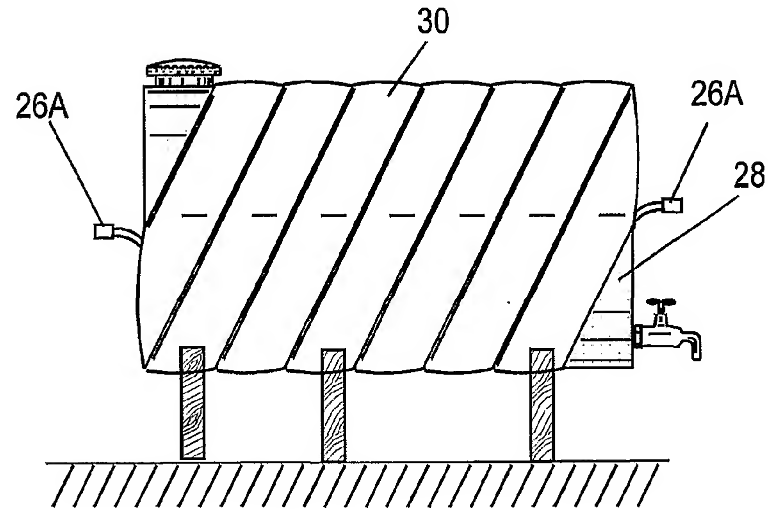


FIGURE 9

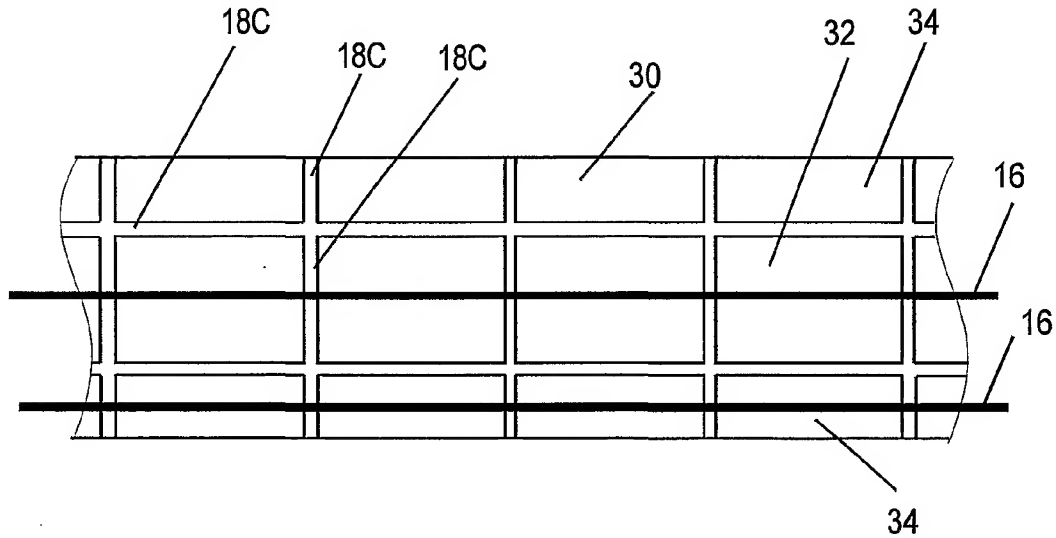


FIGURE 10

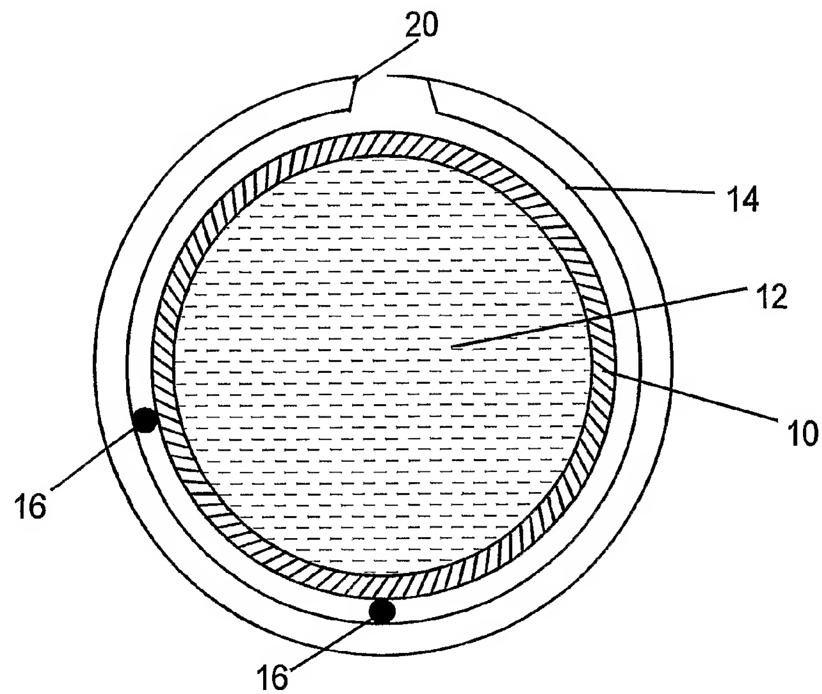


FIGURE 11

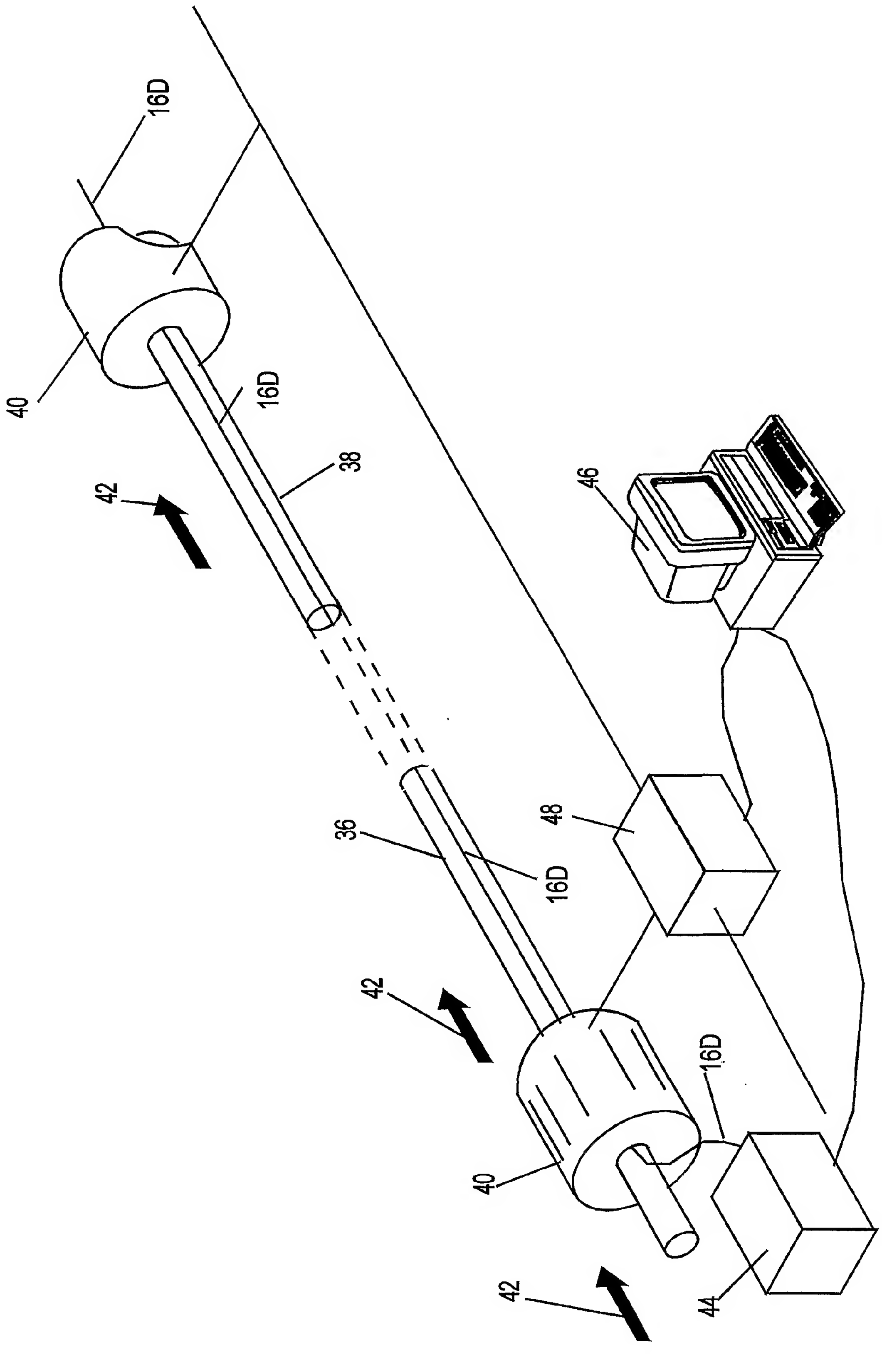


FIGURE 12

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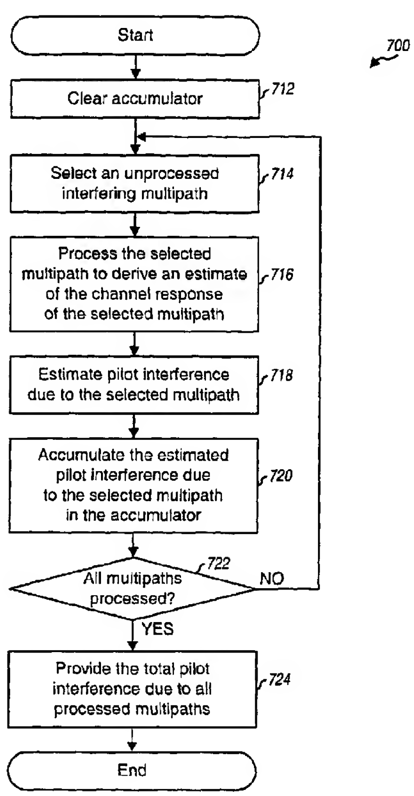
*[Continued on next page]*

(54) Title: METHOD AND APPARATUS FOR CANCELING PILOT INTERFERENCE IN A WIRELESS COMMUNICATION SYSTEM




(57) Abstract: Techniques for canceling pilot interference in a wireless (e.g., CDMA) communication system. In one method, a received signal comprised of a number of signal instances, each including a pilot, is initially processed to provide data samples. Each signal instance's pilot interference may be estimated by despreading the data samples with a spreading sequence for the signal instance, channelizing the despread data to provide pilot symbols, filtering the pilot symbols to estimate the channel response of the signal instance, and multiplying the estimated channel response with the spreading sequence to provide the estimated plot interference. The pilot interference estimates due to all interfering multipaths are combined to derive the total pilot interference, which is subtracted from the data samples to provide pilot-canceled data samples. These samples are then processed to derive demodulated data for each of at least one (desired) signal instance in the received signal.

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# METHOD AND APPARATUS FOR CANCELING PILOT INTERFERENCE IN A WIRELESS COMMUNICATION SYSTEM

**[0001]** This application claims the benefit of provisional U.S. Application Serial No. 60/296,259, entitled "METHOD AND APPARATUS FOR CANCELLATION OF MULTIPLE PILOT SIGNALS," filed June 6, 2001, which is incorporated herein by reference in its entirety for all purposes.

## BACKGROUND

### Field

**[0002]** The present invention relates generally to data communication, and more specifically to techniques for canceling interference due to pilots in a wireless (e.g., CDMA) communication system.

### Background

**[0003]** Wireless communication systems are widely deployed to provide various types of communication such as voice, packet data, and so on. These systems may be based on code division multiple access (CDMA), time division multiple access (TDMA), or some other multiple access technique. CDMA systems may provide certain advantages over other types of systems, including increased system capacity. A CDMA system is typically designed to implement one or more standards, such as IS-95, cdma2000, IS-856, W-CDMA, and TS-CDMA standards, all of which are known in the art.

**[0004]** In some wireless (e.g., CDMA) communication systems, a pilot may be transmitted from a transmitter unit (e.g., a terminal) to a receiver unit (e.g., a base station) to assist the receiver unit perform a number of functions. For example, the pilot may be used at the receiver unit for synchronization with the timing and frequency of the transmitter unit, estimation of the channel response and the quality of the communication channel, coherent demodulation of data transmission, and so on. The pilot is typically generated based on a known data pattern (e.g., a sequence

of all zeros) and using a known signal processing scheme (e.g., channelized with a particular channelization code and spread with a known spreading sequence).

[0005] On the reverse link in a cdma2000 system, the spreading sequence for each terminal is generated based on (1) a complex pseudo-random noise (PN) sequence common to all terminals and (2) a scrambling sequence specific to the terminal. In this way, the pilots from different terminals may be identified by their different spreading sequences. On the forward link in cdma2000 and IS-95 systems, each base station is assigned a specific offset of the PN sequence. In this way, the pilots from different base stations may be identified by their different assigned PN offsets.

[0006] At the receiver unit, a rake receiver is often used to recover the transmitted pilot, signaling, and traffic data from all transmitter units that have established communication with the receiver unit. A signal transmitted from a particular transmitter unit may be received at the receiver unit via multiple signal paths, and each received signal instance (or multipath) of sufficient strength may be individually demodulated by the rake receiver. Each such multipath is processed in a manner complementary to that performed at the transmitter unit to recover the data and pilot received via this multipath. The recovered pilot has an amplitude and phase determined by, and indicative of, the channel response for the multipath. The pilot is typically used for coherent demodulation of various types of data transmitted along with the pilot, which are similarly distorted by the channel response. For each transmitter unit, the pilots for a number of multipaths for the transmitter unit are also used to combine demodulated symbols derived from these multipaths to obtain combined symbols having improved quality.

[0007] On the reverse link, the pilot from each transmitting terminal acts as interference to the signals from all other terminals. For each terminal, the aggregate interference due to the pilots transmitted by all other terminals may be a large percentage of the total interference experienced by this terminal. This pilot interference can degrade performance (e.g., a higher packet error rate) and further reduce the reverse link capacity.

[0008] There is therefore a need for techniques to cancel interference due to pilots in a wireless (e.g., CDMA) communication system.

## SUMMARY

[0009] Aspects of the present invention provide techniques for estimating and canceling pilot interference in a wireless (e.g., CDMA) communication system. A received signal typically includes a number of signal instances (i.e., multipaths). For each multipath to be demodulated (i.e., each desired multipath), the pilots in all multipaths are interference to the data in the desired multipath. If the pilot is generated based on a known data pattern (e.g., a sequence of all zeros) and channelized with a known channelization code (e.g., a Walsh code of zero), then the pilot in an interfering multipath may be estimated as simply a spreading sequence with a phase corresponding to the arrival time of that multipath at the receiver unit. The pilot interference from each interfering multipath may be estimated based on the spreading sequence and an estimate of the channel response of that multipath (which may be estimated based on the pilot). The total pilot interference due to a number of interfering multipaths may be derived and subtracted from the received signal to provide a pilot-canceled signal having the pilot interference removed.

[0010] In one specific embodiment, a method for canceling pilot interference at a receiver unit (e.g., a base station) in a wireless (e.g., cdma2000) communication system is provided. In accordance with the method, a received signal comprised of a number of signal instances, each of which includes a pilot, is initially processed to provide data samples. The data samples are then processed to derive an estimate of the pilot interference due to each of one or more (interfering) signal instances, and the pilot interference estimates are further combined to derive the total pilot interference. The total pilot interference is then subtracted from the data samples to provide pilot-canceled data samples, which are further processed to derive demodulated data for each of at least one (desired) signal instance in the received signal.

[0011] The pilot interference due to each interfering signal instance may be estimated by (1) despread the data samples with a spreading sequence for the signal instance, (2) channelizing the despread samples with a pilot channelization code to provide pilot symbols, (3) filtering the pilot symbols to provide an estimated channel

response of the signal instance, and (4) multiplying the spreading sequence for the signal instance with the estimated channel response to provide the estimated pilot interference. The data demodulation for each desired multipath may be performed by (1) despreading the pilot-canceled data samples with the spreading sequence for the signal instance, (2) channelizing the despread samples with a data channelization code to provide data symbols, and (3) demodulating the data symbols to provide the demodulated data for the signal instance. For improved performance, the pilot estimation and cancellation may be performed at a sample rate that is higher than the PN chip rate.

**[0012]** Various aspects, embodiments, and features of the invention are described in further detail below.

## BRIEF DESCRIPTION OF THE DRAWINGS

**[0013]** The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

**[0014]** FIG. 1 is a diagram of a wireless communication system;

**[0015]** FIG. 2 is a simplified block diagram of an embodiment of a base station and a terminal;

**[0016]** FIG. 3 is a block diagram of an embodiment of a modulator for the reverse link in cdma2000;

**[0017]** FIG. 4 is a block diagram of an embodiment of a rake receiver;

**[0018]** FIG. 5 is a block diagram of a specific embodiment of a finger processor within the rake receiver, which is capable of estimating and canceling pilot interference in addition to performing data demodulation;

**[0019]** FIGS. 6A and 6B are diagrams that graphically illustrate the processing of the data samples to derive estimates of pilot interference, in accordance with a specific implementation of the invention;

[0020] FIG. 7 is a flow diagram of an embodiment of a process to derive the total pilot interference for a number of multipaths; and

[0021] FIG. 8 is a flow diagram of an embodiment of a process to data demodulate a number of multipaths with pilot interference cancellation.

## DETAILED DESCRIPTION

[0022] FIG. 1 is a diagram of a wireless communication system 100 that supports a number of users and wherein various aspects and embodiments of the invention may be implemented. System 100 provides communication for a number of cells, with each cell being serviced by a corresponding base station 104. A base station is also commonly referred to as a base-station transceiver system (BTS), an access point, or a Node B. Various terminals 106 are dispersed throughout the system. Each terminal 106 may communicate with one or more base stations 104 on the forward and reverse links at any given moment, depending on whether or not the terminal is active and whether or not it is in soft handoff. The forward link (i.e., downlink) refers to transmission from the base station to the terminal, and the reverse link (i.e., uplink) refers to transmission from the terminal to the base station.

[0023] A signal transmitted from a terminal may reach a base station via one or multiple signal paths. These signal paths may include a straight path (e.g., signal path 110a) and reflected paths (e.g., signal path 110b). A reflected path is created when the transmitted signal is reflected off a reflection source and arrives at the base station via a different path than the line-of-sight path. The reflection sources are typically artifacts in the environment in which the terminal is operating (e.g., buildings, trees, or some other structures). The signal received by each antenna at the base station may thus comprise a number of signal instances (or multipaths) from one or more terminals.

[0024] In system 100, a system controller 102 (which is also often referred to as a base station controller (BSC)) couples to base stations 104, provides coordination and control for the base stations coupled to it, and further controls the routing of calls to terminals 106 via the coupled base stations. System controller 102 may further couple to a public switched telephone network (PSTN) via a mobile switching center

(MSC), and to a packet data network via a packet data serving node (PDSN), which are not shown in FIG. 1. System 100 may be designed to support one or more CDMA standards such as cdma2000, IS-95, IS-856, W-CDMA, TS-CDMA, some other CDMA standards, or a combination thereof. These CDMA standards are known in the art and incorporated herein by reference.

**[0025]** Various aspects and embodiments of the invention may be applied for the forward and reverse links in various wireless communication systems. For clarity, the pilot interference cancellation techniques are specifically described for the reverse link in a cdma2000 system.

**[0026]** FIG. 2 is a simplified block diagram of an embodiment of base station 104 and terminal 106. On the reverse link, at terminal 106, a transmit (TX) data processor 214 receives various types of "traffic" such as user-specific data from a data source 212, messages, and so on. TX data processor 214 then formats and codes the different types of traffic based on one or more coding schemes to provide coded data. Each coding scheme may include any combination of cyclic redundancy check (CRC), convolutional, Turbo, block, and other coding, or no coding at all. Interleaving is commonly applied when error correcting codes are used to combat fading. Other coding scheme may include automatic repeat request (ARQ), hybrid ARQ, and incremental redundancy repeat. Typically, different types of traffic are coded using different coding schemes. A modulator (MOD) 216 then receives pilot data and the coded data from TX data processor 214, and further processes the received data to generate modulated data.

**[0027]** FIG. 3 is a block diagram of an embodiment of a modulator 216a, which may be used for modulator 216 in FIG. 2. For the reverse link in cdma2000, the processing by modulator 216a includes covering the data for each of a number of code channels (e.g., traffic, sync, paging, and pilot channels) with a respective Walsh code, $C_{chx}$, by a multiplier 312 to channelize the user-specific data (packet data), messages (control data), and pilot data onto their respective code channels. The channelized data for each code channel may be scaled with a respective gain, $G_i$, by a unit 314 to control the relative transmit power of the code channels. The scaled data for all code channels for the inphase (I) path is then summed by a summer 316a to

provide I-channel data, and the scaled data for all code channels for the quadrature (Q) path is summed by a summer 316b to provide Q-channel data.

**[0028]** FIG. 3 also shows an embodiment of a spreading sequence generator 320 for the reverse link in cdma2000. Within generator 320, a long code generator 322 receives a long code mask assigned to the terminal and generates a long pseudo-random noise (PN) sequence with a phase determined by the long code mask. The long PN sequence is then multiplied with an I-channel PN sequence by a multiplier 326a to generate an I spreading sequence. The long PN sequence is also delayed by a delay element 324, multiplied with a Q-channel PN sequence by a multiplier 326b, decimated by a factor of two by element 328, and covered with a Walsh code ($C_s$ = +-) and further spread with the I spreading sequence by a multiplier 330 to generate a Q spreading sequence. The I-channel and Q-channel PN sequences form the complex short PN sequence used by all terminals. The I and Q spreading sequences form the complex spreading sequence, $S_k$, that is specific to the terminal.

**[0029]** Within modulator 216a, the I-channel data and the Q-channel data ($D_{chI} + jD_{chQ}$) are spread with the I and Q spreading sequences ($S_{kI} + jS_{kQ}$), via a complex multiply operation performed by a multiplier 340, to generate I spread data and Q spread data ($D_{spI} + jD_{spQ}$). The complex despreading operation may be expressed as:

**[0030]**

$$\begin{aligned} D_{spI} + jD_{spQ} &= (D_{chI} + jD_{chQ}) \cdot (S_{kI} + jS_{kQ}) \ , \\ &= (D_{chI}S_{kI} - D_{chQ}S_{kQ}) + j(D_{chI}S_{kQ} + D_{chQ}S_{kI}) \ . \end{aligned}$$ Eq (1)

**[0031]** The I and Q spread data comprises the modulated data provided by modulator 216a.

**[0032]** The modulated data is then provided to a transmitter (TMTR) 218a and conditioned. Transmitter 218a is an embodiment of transmitter 218 in FIG. 2. The signal conditioning includes filtering the I and Q spread data with filters 352a and 352b, respectively, and upconverting the filtered I and Q data with $\cos(w_c t)$ and $\sin(w_c t)$, respectively, by multipliers 354a and 354b. The I and Q components from

multipliers 354a and 354b are then summed by a summer 356 and further amplified with a gain, $G_o$, by a multiplier 358 to generate a reverse link modulated signal.

**[0033]** Referring back to FIG. 2, the reverse link modulated signal is then transmitted via an antenna 220 and over a wireless communication link to one or more base stations.

**[0034]** At base station 104, the reverse link modulated signals from a number of terminals are received by each of one or more antennas 250. Multiple antennas 250 may be used to provide spatial diversity against deleterious path effect such as fading. As an example, for a base station that supports three sectors, two antennas may be used for each sector and the base station may then include six antennas. Any number of antennas may thus be employed at the base station.

**[0035]** Each received signal is provided to a respective receiver (RCVR) 252, which conditions (e.g., filters, amplifies, downconverts) and digitizes the received signal to provide data samples for that received signal. Each receive signal may include one or more signal instances (i.e., multipaths) for each of a number of terminals.

**[0036]** A demodulator (DEMOD) 254 then receives and processes the data samples for all received signals to provide recovered symbols. For cdma2000, the processing by demodulator 254 to recover a data transmission from a particular terminal includes (1) despreading the data samples with the same spreading sequence used to spread the data at the terminal, (2) channelizing the despread samples to isolate or channelize the received data and pilot onto their respective code channels, and (3) coherently demodulating the channelized data with a recovered pilot to provide demodulated data. Demodulator 254 may implement a rake receiver that can process multiple signal instances for each of a number of terminals, as described below.

**[0037]** A receive (RX) data processor 256 then receives and decodes the demodulated data for each terminal to recover the user-specific data and messages transmitted by the terminal on the reverse link. The processing by demodulator 254

and RX data processor 256 is complementary to that performed by modulator 214 and TX data processor 212, respectively, at the terminal.

[0038] FIG. 4 is a block diagram of an embodiment of a rake receiver 254a, which is capable of receiving and demodulating the reverse link modulated signals from a number of terminals 106. Rake receiver 254a includes one or more (L) sample buffers 408, one or more (M) finger processors 410, a searcher 412, and a symbol combiner 420. The embodiment in FIG. 4 shows all finger processor 410 coupled to the same symbol combiner 420.

[0039] Due to the multipath environment, the reverse link modulated signal transmitted from each terminal 106 may arrive at base station 104 via a number of signal paths (as shown in FIG. 1), and the received signal for each base station antenna typically comprises a combination of different instances of the reverse link modulated signal from each of a number of terminals. Each signal instance (or multipath) in a received signal is typically associated with a particular magnitude, phase, and arrival time (i.e., a time delay or time offset relative to CDMA system time). If the difference between the arrival times of the multipaths is more than one PN chip at the base station, then each received signal, $y_l(t)$, at the input to a respective receiver 252 may be expressed as:

[0040] $$y_l(t) = \sum_j \sum_i p_{i,j,l}(t) x_j(t - \hat{t}_{i,j,l}) \quad + n(t) \quad ,$$ Eq

(2)

[0041] where

[0042] $x_j(t)$ is the $j$-th reverse link modulated signal transmitted by the $j$-th terminal;

[0043] $\hat{t}_{i,j,l}$ is the arrival time, at the $l$-th antenna, of the $i$-th multipath relative to the time the $j$-th reverse link modulated signal, $x_j(t)$, is transmitted;

[0044] $p_{i,j,l}(t)$ represents the channel gain and phase for the $i$-th multipath for the $j$-th terminal at the $l$-th antenna, and is a function of the fading process;

[0045] $\sum_j$ is the summation for all reverse link modulated signals in the $l$-th received signal;

[0046] $\sum_i$ is the summation for all multipaths of each reverse link modulated signal in the $l$-th received signal; and

[0047] $n(t)$ represents the real-valued channel noise at RF plus internal receiver noise.

[0048] Each receiver 252 amplifies and frequency downconverts a respective received signal, $y_l(t)$, and further filters the signal with a received filter that is typically matched to the transmit filter (e.g., filter 352) used at the terminal to provide a conditioned signal. Each receiver unit 252 then digitizes the conditioned signal to provide a respective stream of data samples, which is then provided to a respective sample buffer 408.

[0049] Each sample buffer 408 stores the received data samples and further provides the proper data samples to the appropriate processing units (e.g., finger processors 410 and/or searcher 412) at the appropriate time. In one design, each buffer 408 provides the data samples to a respective set of finger processors assigned to process the multipaths in the received signal associated with the buffer. In another design, a number of buffers 408 provide data samples (e.g., in a time division multiplexed manner) to a particular finger processor that has the capability to process a number of multipaths in a time division multiplexed manner. Sample buffers 408a through 408$l$ may also be implemented as a single buffer of the appropriate size and speed.

[0050] Searcher 412 is used to search for strong multipaths in the received signals and to provide an indication of the strength and timing of each found multipath that meets a set of criteria. The search for multipaths of a particular terminal is typically performed by correlating the data samples for each received signal with the terminal's spreading sequence, locally generated at various chip or sub-chip offsets (or phases). Due to the pseudo-random nature of the spreading

sequence, the correlation of the data samples with the spreading sequence should be low, except when the phase of the locally-generated spreading sequence is time-aligned with that of a multipath, in which case the correlation results in a high value.

**[0051]** For each reverse link modulated signal, $x_j(t)$, searcher 412 may provide a set of one or more time offsets, $t_{i,j,l}$, for a set of one or more multipaths found for that reverse link modulated signal (possibly along with the signal strength of each found multipath). The time offsets, $t_{i,j,l}$, provided by searcher 412 are relative to the base station timing or CDMA system time, and are related to the time offsets, $\hat{t}_{i,j,l}$, shown in equation (2) which are relative the time of signal transmission.

**[0052]** Searcher 412 may be designed with one or multiple searcher units, each of which may be designed to search for multipaths over a respective search window. Each search window includes a range of spreading sequence phases to be searched. The searcher units may be operated in parallel to speed up the search operation. Additionally or alternatively, searcher 412 may be operated at a high clock rate to speed up the search operation. Searcher and searching are described in further detail in U.S. Patent Nos. 5,805,648, 5,781,543, 5,764,687, and 5,644,591, all of which are incorporated herein by reference.

**[0053]** Each finger processor 410 may then be assigned to process a respective set of one or more multipaths of interest (e.g., multipaths of sufficient strength, as determined by controller 260 based on the signal strength information provided by searcher 412). Each finger processor 410 then receives, for each assigned multipath, the following: (1) the data samples for the received signal that includes the assigned multipath, (2) either the time offset, $t_{i,j,l}$, of the assigned multipath or a spreading sequence, $S_{i,j,l}$, with a phase corresponding to the time offset, $t_{i,j,l}$ (which may be generated by a spreading sequence generator 414), and (3) the channelization code (e.g., the Walsh code) for the code channel to be recovered. Each finger processor 410 then processes the received data samples and provides demodulated data for each assigned multipath. The processing by finger processor 410 is described in further detail below.

**[0054]** Symbol combiner 420 receives and combines the demodulated data (i.e., the demodulated symbols) for each terminal. In particular, symbol combiner 420 receives the demodulated symbols for all assigned multipaths for each terminal and, depending on the design of the finger processors, may time-align (or deskew) the symbols to account for differences in the time offsets for the assigned multipaths. Symbol combiner 420 then combines the time-aligned demodulated symbols for each terminal to provide recovered symbols for the terminal. Multiple symbol combiners may be provided to concurrently combine symbols for multiple terminals. The recovered symbols for each terminal are then provided to RX data processor 256 and decoded.

**[0055]** The processing of the multipaths may be performed based on various demodulator designs. In a first demodulator design, one finger processor is assigned to process a number of multipaths in a received signal. For this design, the data samples from the sample buffer may be processed in "segments" covering a particular time duration (i.e., a particular number of PN chips) and starting at some defined time boundaries. In a second demodulator design, multiple finger processors are assigned to process multiple multipaths in the received signal. Various aspects and embodiments of the invention are described for the first demodulator design.

**[0056]** The pilot interference cancellation may also be performed based on various schemes. In a first pilot interference cancellation scheme that is based on the first demodulator design, the channel response of a particular multipath is estimated based on a segment of data samples, and the estimated channel response is then used to derive an estimate of the pilot interference due to this multipath for the same segment. This scheme may provide improved pilot interference cancellation. However, this scheme also introduces additional processing delays in the data demodulation for the multipath since the segment of data samples is first processed to estimate and cancel the pilot interference before the data demodulation can proceed on the same segment.

**[0057]** In a second pilot interference cancellation scheme that is also based on the first demodulator design, the channel response of a particular multipath is estimated based on a segment of data samples, and the estimated channel response is

then used to derive an estimate of the pilot interference due to this multipath for the next segment. This scheme may be used to reduce (or possibly eliminate) the additional processing delays in the data demodulation resulting from the pilot interference estimation and cancellation. However, since the link conditions may continually change over time, the time delay between the current and next segments should be kept sufficiently short such that the channel response estimate for the current segment is still accurate in the next segment. For clarity, the pilot interference estimation and cancellation are described below for the second scheme.

**[0058]** FIG. 5 is a block diagram of a specific embodiment of a finger processor 410x, which is capable of estimating and canceling pilot interference in addition to performing the data demodulation. Finger processor 410x may be used for each finger processor 410 in rake receiver 254a shown in FIG. 4. In the following description, FIG. 5 shows the processing elements and FIGS. 6A and 6B graphically show the timing for the pilot interference estimation and cancellation.

**[0059]** Finger processor 410x is assigned to demodulate one or more "desired" multipaths in a particular received signal. Sample buffer 408x stores data samples for the received signal that includes the multipaths assigned to finger processor 410x. Buffer 408x then provides the appropriate data samples (in segments) to the finger processor when and as they are needed. In the embodiment shown in FIG. 5, finger processor 410x includes a resampler 522, a pilot estimator 520 (or channel estimator), a summer 542, a data demodulation unit 550, and a pilot interference estimator 530.

**[0060]** For each desired multipath to be demodulated by finger processor 410x, the data in all other multipaths and the pilots in all multipaths in the same received signal act as interference to this multipath. Since the pilot is generated based on a known data pattern (e.g., typically a sequence of all zeros) and processed in a known manner, the pilots in the "interfering" multipaths may be estimated and removed from the desired multipath to improve the signal quality of the data component in the desired multipath. Finger processor 410x is capable of estimating and canceling the pilot interference due to a number of multipaths found in the received signal, including the pilot of the desired multipath, as described below.

[0061] In an embodiment, the pilot interference estimation and cancellation and the data demodulation are performed in "bursts". For each burst (i.e., each processing cycle), a segment of data samples for a particular number PN chips are processed to estimate the pilot interference due to a particular multipath. In a specific embodiment, each segment comprises data samples for one symbol period, which may be 64 PN chips for cdma2000. However, other segment sizes may also be used (e.g., for data symbols of other durations), and this is within the scope of the invention. As described below, the data demodulation may be performed in parallel and in a pipelined manner with the pilot interference estimation to increase processing throughput and possibly reduce the overall processing time.

[0062] To derive an estimate of the pilot interference due to the $m$-th multipath (where $m = (i, j, l)$ and is the notation for the $i$-th multipath for the $j$-th reverse link modulated signal found in the $l$-th received signal), a segment of data samples is initially provided from buffer 408x to a resampler 522 within finger processor 410x. Resampler 522 may then perform decimation, interpolation, or a combination thereof, to provide decimated data samples at the chip rate and with the proper "fine-grain" timing phase.

[0063] FIG. 6A graphically illustrates an embodiment of the resampling performed by resampler 522. The received signal is typically oversampled at a sample rate that is multiple (e.g., 2, 4, or 8) times the chip rate to provide higher time resolution. The data samples are stored to sample buffer 408x, which thereafter provides a segment of (e.g., 512) data samples for each processing cycle. Resampler 522 then "resamples" the data samples received from buffer 408x to provide samples at the chip rate and with the proper timing phase.

[0064] As shown in FIG. 6A, if the received signal is sufficiently oversampled (e.g., at 8 times the chip rate), then the resampling for the $m$-multipath may be performed by providing every, e.g., 8-th data sample received from the buffer, with the selected data samples being the ones most closely aligned to the timing of the peak of the $m$-th multipath. The $m$-th multipath is typically a multipath assigned for data demodulation, and the multipath's time offset, $t_m$, may be determined and provided by searcher 412. However, pilot interference due to multipaths that are not

assigned for data demodulation may also be estimated and canceled, so long as the time offset of each such multipath is known. Each multipath's time offset, $t_m$, may be viewed as comprising an integer number of symbol periods and a fractional portion of a symbol period (i.e., $t_m = t_{full,m} + t_{frac,m}$) relative to the base station timing or CDMA system time, where a symbol period is determined by the length of the channelization code (e.g., 64 PN chips for cdma2000). The fractional part of the time offset, $t_{frac,m}$, may be used to select the particular segment of data samples to provide to resampler 522 and for decimation. In the example shown in FIG. 6A, the fractional part of the time offset for the $m$-th multipath is $t_{frac,m} = 5$, data sample segment 622 is provided by buffer 408x, and the decimated data samples provided by resampler 522 are represented by the shaded boxes.

[0065] For some other receiver design in which the received signal is not sufficiently oversampled, then interpolation may alternatively or additionally be performed along with decimation to derive new samples at the proper timing phase, as is known in the art.

[0066] Within pilot estimator 520, a despreader 524 receives the decimated data samples and a (complex-conjugate) spreading sequence, $S_m^*(k)$, having a phase corresponding to the time offset, $t_m$, of the $m$-th multipath whose pilot interference is to be estimated. The spreading sequence, $S_m^*(k)$, may be provided by spreading sequence generator 414. For the reverse link in cdma2000, the spreading sequence, $S_m^*(k)$, may be generated as shown for spreading sequence generator 320 in FIG. 3. And as shown in FIG. 6A, a segment of the spreading sequence, $S_m^*(k)$, of the same length and with the same timing phase as the data sample segment is used for the despreading (i.e., the spreading sequence, $S_m^*(k)$ is time-aligned with the decimated data samples).

[0067] Despreader 524 (which may be implemented as a complex multiplier such as multiplier 340 shown in FIG. 3) despreads the decimated data samples with the spreading sequence, $S_m^*(k)$, and provides despread samples. A pilot channelizer

526 then multiplies the despread samples with the channelization code, $C_{pilot,m}$, used for the pilot at the terminal (e.g., a Walsh code of zero for cdma2000). The decovered pilot samples are then accumulated over a particular accumulation time interval to provide pilot symbols. The accumulation time interval is typically an integer multiple of the pilot channelization code length. If the pilot data is covered with a channelization code of zero (as in cdma2000), then the multiplication with the channelization code, $C_{pilot,m}$, may be omitted and pilot channelizer 526 simply performs the accumulation of the despread samples from despreader 524. In a specific embodiment, one pilot symbol is provided for each segment, which has a size of one symbol period.

**[0068]** The pilot symbols from pilot channelizer 526 are then provided to a pilot filter 528 and filtered based on a particular lowpass filter response to remove noise. Pilot filter 528 may be implemented as a finite impulse response filter (FIR), an infinite impulse response (IIR) filter, or some other filter structure. Pilot filter 528 provides pilot estimates, $P_m(k)$, which are indicative of the channel response (i.e., the gain and phase, $a_m \cdot e^{j\theta_m}$) of the $m$-th multipath. Each pilot estimate, $P_m(k)$, is thus a complex value. The pilot estimates are provided at sufficient rate such that non-insignificant changes in the channel response of the multipath are captured and reported. In a specific embodiment, one pilot estimate is provided for each segment, which has a size of one symbol.

**[0069]** Pilot interference estimator 530 then estimates the pilot interference due to the $m$-th multipath for the next segment. To estimate the pilot interference, the pilot data and the pilot channelization code, $C_{pilot,m}$, for the $m$-th multipath are provided to a pilot channelizer 532, which channelizes the pilot data with the pilot channelization code to provide channelized pilot data. A spreader 534 then receives and spreads the channelized pilot data with a spreading sequence, $S_m(k+\mathrm{N})$, to generate spread pilot data (i.e., processed pilot data). The spreading sequence, $S_m(k+\mathrm{N})$, has a phase corresponding to the time offset, $t_m$, of the $m$-th interfering multipath and is further advanced by N PN chips for the next segment, as shown in FIG. 6A. If the pilot data is a sequence of all zeros and the pilot channelization code

is also a sequence of all zeros (as in cdma2000), then pilot channelizer 532 and spreader 534 may be omitted and the spread pilot data is simply the spreading sequence, $S_m(k+\mathrm{N})$.

**[0070]** A multiplier 536 then receives and multiplies the spread pilot data with the pilot estimates, $P_m(k)$, from pilot filter 528 to provide an estimate of the pilot interference, $I_{pilot,m}(k+\mathrm{N})$, due to the *m*-th multipath for the next segment. Since the pilot estimates, $P_m(k)$, are derived from the current segment and used to derive the estimated pilot interference for the next segment, prediction techniques may be used to derive pilot predictions for the next segment based on the pilot estimates. These pilot predictions may then be used to derive the estimated pilot interference for the next segment.

**[0071]** In an embodiment, multiplier 536 provides the estimated pilot interference due to the *m*-th multipath at the sample rate (e.g., 8x the chip rate) and with the timing phase of the *m*-th multipath. This allows the estimated pilot interferences for all multipaths (which have different time offsets that are typically not all aligned to the PN chip timing boundaries) to be accumulated at a higher time resolution. The estimated pilot interference, $I_{pilot,m}(k+\mathrm{N})$, for the *m*-th multipath, which includes the same number of interference samples as for the data sample segment, is then provided to an interference accumulator 538. As shown in FIG. 6A, the interference samples for the *m*-th multipath are stored (or accumulated with the interference samples already stored) at locations in the accumulator determined by the fractional part of the multipath's time offset.

**[0072]** To derive the total pilot interference for all multipaths in a given received signal, the processing described above may be iterated a number of times, one iteration or processing cycle for each interfering multipath for which the pilot interference is to be estimated and canceled from a desired multipath. The pilot interference cancellation is typically performed for the multipaths received via the same antenna, not cross antennas, because the channel estimate from one antenna is typically not good for another antenna. If the same finger processor hardware is used for multiple iterations, then the processing may be performed in bursts, with each

burst being performed on a respective segment of data samples determined by the multipath's fractional time offset.

**[0073]** Prior to the first iteration, accumulator 538 is cleared or reset. For each iteration, the estimated pilot interference, $I_{pilot,m}$, due to the current multipath is accumulated with the accumulated pilot interference for all prior-processed multipaths. However, as shown in FIG. 6A, the estimated pilot interference, $I_{pilot,m}$, is accumulated with samples in a specific section of accumulator 538, which is determined by the current multipath's time offset. After all interfering multipaths have been processed, the accumulated pilot interference in accumulator 538 comprises the total pilot interference, $I_{pilot}$, due to all processed multipaths.

**[0074]** FIG. 6A also shows an embodiment of accumulator 538. While finger processor 410x performs data demodulation for the $m$-th multipath for the current segment (using the total pilot interference, $I_{pilot}(k)$, derived earlier and stored in one section of accumulator 538), the pilot interference due to the $m$-th multipath, $I_{pilot,m}(k+\mathrm{N})$, for the next segment may be estimated and accumulated in another section of the accumulator.

**[0075]** The pilot for the $m$-th multipath is interference to all multipaths in the received signal, including the $m$-th multipath itself. For a demodulator design in which the multiple finger processors are assigned to process a number of multipaths in a received signal for a given terminal, the estimated pilot interference, $I_{pilot,m}$, due to the $m$-th multipath may be provided to other finger processors assigned to process other multipaths in the same received signal.

**[0076]** For the demodulation to recover the data on the $m$-th multipath, the data samples for a segment are provided from buffer 408x to resampler 522. Resampler 522 then resamples the received data samples to provide decimated data samples at the chip rate and with the proper timing phase for this multipath. The decimated data samples are processed as described above to provide the pilot estimates, $P_m(k)$.

[0077] Correspondingly, interference samples for the total pilot interference, $I_{pilot}(k)$, for the same segment are provided from accumulator 538 to a resampler 540. Resampler 540 similarly resamples the received interference samples to provide decimated interference samples at the chip rate and with the proper timing phase for the $m$-th multipath. Summer 542 then receives and subtracts the decimated interference samples from the decimated data samples to provide pilot-canceled data samples.

[0078] Within data demodulation unit 550, a despreader 544 receives and despreads the pilot-canceled data samples with a (complex-conjugate) spreading sequence, $S_m^*(k)$, to provide despread samples. The spreading sequence, $S_m^*(k)$, has a phase corresponding to the time offset, $t_m$, of the $m$-th multipath. A data channelizer 546 then multiplies the despread samples with the channelization code, $C_{ch,m}$, used for the code channel being recovered by the finger processor. The channelized data samples are then accumulated over the length of the channelization code, $C_{ch,m}$, to provide data symbols.

[0079] A data demodulator 548 then receives and demodulates the data symbols with the pilot estimates, $P_m(k)$, to provide demodulated symbols (i.e., demodulated data) for the $m$-th multipath, which are then provided to symbol combiner 420. The data demodulation and symbol combining may be achieved as described in the aforementioned U.S Patent No. 5,764,687 patent. The '687 patent describes BPSK data demodulation for IS-95 by performing dot product between the despread data and the filtered pilot. The demodulation of QPSK modulation, which is used in cdma2000 and W-CDMA, is a straight-forward extension of the techniques described in the '687 patent. That is, instead of dot product, both dot product and cross-product are used to recover the inphase and quadrature streams.

[0080] As noted above, the data demodulation for the $m$-th multipath may be performed in parallel and in a pipelined manner with the pilot interference estimation. While despreader 544 and data channelizer 546 are processing the pilot-canceled data samples for the current segment (with the spreading sequence, $S_m^*(k)$, and the

channelization code, $C_{ch,m}$) to provide the data symbols for the $m$-th multipath, despreader 524 and pilot channelizer 526 may process the same data samples for the current segment (with the spreading sequence, $S_m^*(k)$, and the pilot channelization code, $C_{pilot,m}$) to provide the pilot symbols for this multipath. The pilot symbols are filtered by pilot filter 528 to provide pilot estimates, $P_m(k)$, for the multipath. Pilot interference estimator 530 then derives the estimated pilot interference, $I_{pilot,m}(k+\mathrm{N})$, due to this multipath for the following segment, as described above. In this manner, while data demodulation is performed on the current segment using the total pilot interference, $I_{pilot}(k)$, derived from a prior segment, pilot interference for the next segment is estimated and stored to another section of the accumulator, to be used for the next segment.

**[0081]** In an embodiment, the pilot for a particular multipath being demodulated is estimated based on the "raw" received data samples (from sample buffer 408x) as described above, and not based on the pilot-canceled data samples (from accumulator 538). In another embodiment, the pilot may be estimated based on the pilot-canceled data samples if the total pilot interference includes some or all of the interfering pilots except for the pilot of the multipath being demodulated (i.e., the pilot of the multipath being demodulated is included in the pilot-canceled data samples). This alternative embodiment may provide an improved estimate of the channel response of the multipath being demodulated, and is especially advantageous for the reverse link where the pilot estimation is typically the limiting factor in dealing with a weak multipath. The same "other pilots canceled" data samples that is used for pilot estimation may also be processed to recover the data for the multipath, which is advantageous for a finger processor architecture that performs both pilot estimation and data demodulation in parallel on the same data sample stream. The same concept may be used to estimate the channel response of a particular interfering multipath (i.e., the estimated channel response may be based on either the raw data samples or the "other pilots canceled" data samples having interfering pilots except for the pilot of that particular multipath removed).

[0082] FIGS. 6A and 6B are diagrams that illustrate the processing of the data samples to derive estimates of pilot interference, in accordance with a specific implementation of the invention. In the example shown in FIGS. 6A and 6B, the received signal includes three multipaths that are associated with time offsets of $t_1$, $t_2$, and $t_3$. The received signal is digitized at a sample rate that is 8 times the chip rate to provide data samples, which are stored to the sample buffer. These multipaths may or may not be sampled at their peaks.

[0083] In the example shown in FIGS. 6A and 6B, each segment included 512 data samples for a symbol period of 64 PN chips. The pilot interference is estimated for each of the three multipaths and for each symbol period. The symbol timing for each multipath is determined by the multipath's fractional time offset. If the fractional time offsets of the multipaths are not the same, which is generally true, then the symbol timing for these multipaths will be different and will be associated with different data sample segments. In an embodiment, the multipaths are processed in an order based on their fractional time offsets, with the multipath having the smallest fractional time offset being processed first and the multipath having the largest fractional time offset being processed last. This processing order ensures that the total pilot interference is derived and available for each multipath when it is processed.

[0084] In FIG. 6A, for the $n$-th symbol period for the $m$-th multipath with a fractional time offset of $t_{frac,m} = 5$, resampler 522 receives data samples 5 through 516 from the sample buffer and provides to despreader 524 data samples 5, 13, 20, and so on, and 509, which are represented by the shaded boxes. Correspondingly, despreader 524 receives the spreading sequence, $S_m^*(k)$, with a phase corresponding to the same time offset of $t_m$, and despreads the decimated data samples with the spreading sequence. A pilot estimate, $P_m(k)$, is then derived based on the despread samples for this segment, as described above.

[0085] To derive the estimated pilot interference due to the $m$-th multipath, spreader 534 receives the spreading sequence, $S_m(k+\mathrm{N})$, corresponding to the next

segment. Multiplier 536 then multiplies the spreading sequence, $S_m(k+\mathrm{N})$, with the pilot estimate, $P_m(k)$, derived from the current segment to provide the estimated pilot interference, $I_{pilot,m}(k+\mathrm{N})$, for the next segment. The estimated pilot interference, $I_{pilot,m}(k+\mathrm{N})$, comprises interference samples 517 through 1028, which are accumulated with the samples at the same indices 517 through 1028 in the interference accumulator, as shown in FIG. 6. In this way, the fractional time offset of the $m$-th multipath is accounted for in the derivation of the total pilot interference.

**[0086]** For the data demodulation of the $m$-th multipath for the $n$-th symbol period, the same segment of interference samples 5 through 516 are provided from accumulator 538 to resampler 540. Resampler 540 then provides to summer 542 interference samples 5, 13, 20, and so on, and 509 (which are also shown by the shaded boxes), corresponding to the same-indexed data samples provided by resampler 522. The data demodulation of the pilot-canceled data samples is then performed as described above. Each multipath may be processed in similar manner. However, since each multipath may be associated with a different time offset, different decimated data and interference samples may be operated on.

**[0087]** FIG. 6B shows the three data sample segments, the decimated data samples, and the three spreading sequences used to derive the estimated pilot interferences due to the three multipaths.

**[0088]** In another demodulator design, the pilot interference estimation/cancellation and the data demodulation may be performed in real-time (e.g., as data samples are received), if sufficient processing capabilities are provided. For example, M finger processors may be assigned to concurrently process M multipaths in a received signal. For each symbol period, each finger processor can derive a pilot estimate for that symbol period, which is then used to derive the estimated pilot interference due to that finger processor's assigned multipath for the next symbol period. A summer then sums the estimated pilot interferences from all M finger processors (taken into account their respective time offsets), and the total pilot interference for the next symbol period is stored in the interference accumulator.

[0089] The total pilot interference may then be subtracted from the data samples as they are received for the next symbol period, and the same pilot-canceled data samples may be provided to all M finger processors for data demodulation. (These finger processors are also provided with the received data samples, without the pilot cancellation, which are used to derive the pilot estimates.) In this way, the data demodulation may be performed on pilot-canceled data samples in real time, and the sample buffer may possibly be eliminated. For the scheme in which the pilot estimate is used to derive the estimated pilot interference for the same segment (and not the next segment), the data samples may be temporarily stored (e.g., for one symbol period) while the total pilot interference is derived.

[0090] For the demodulator design in which the same data samples are processed multiple times (e.g., if one finger processor is assigned to process a number of multipaths), the sample buffer may be designed and operated in a manner to ensure that the data samples are not inadvertently dropped. In an embodiment, the sample buffer is designed to receive incoming data samples while providing stored data samples to the finger processor(s). This may be achieved by implementing the sample buffer in a manner such that stored data samples may be read from one part of the buffer while new data samples are written into another part of the buffer. The sample buffer may be implemented as a double buffer or multiple buffers, a multi-port buffer, a circular buffer, or some other buffer design. The interference accumulator may be implemented in similar manner as the sample buffer (e.g., as a circular buffer).

[0091] For the above demodulator design, to avoid overwriting samples that are still being processed, the capacity of the sample buffer may be selected to be at least twice the time required to derive the total pilot interference for all M multipaths (with the relationship between time and buffer capacity being defined by the sample rate). If a different data sample segment may be used for each of the M multipaths, then the capacity of the sample buffer may be selected to be at least $(2 \cdot N \cdot N_{os})$ for each received signal assigned to the sample buffer, where N is the duration of data samples used to derive the estimated pilot interference for one multipath and $N_{os}$ is the oversampling factor for the data samples (which is defined as the ratio of the

sample rate over the chip rate). For the above example in which a segment of one symbol period (e.g., N = 64 PN chips) is processed for each multipath, a buffer of two symbol periods would be able to provide a segment of one symbol period of data samples for each multipath regardless of its fractional time offset. And if the oversample rate is $N_{os} = 8$, then the minimum size of the buffer is $(2 \cdot N \cdot N_{os} = 2 \cdot 64 \cdot 8 = 1024)$ data samples.

**[0092]** Similarly, the capacity of the interference accumulator may be selected to be at least $(3 \cdot N \cdot N_{os})$. The extra symbol period for the interference accumulator (i.e., $3 \cdot N$ instead of $2 \cdot N$) is to account for the fact that the estimated pilot interference is derived for the next segment.

**[0093]** As noted above, the estimated pilot interference derived from one data sample segment may be cancelled from a later data sample segment. For a mobile terminal, the communication link and, consequently, the channel response of the various multipaths are constantly changing. Therefore, it is desirable to reduce the delay between the data samples from which the pilot interference is estimated and the data samples from which that estimated pilot interference is canceled. This delay may be as great as $2 \cdot N$ chips.

**[0094]** By selecting a sufficiently small value for N, the channel response of each multipath may be expected to remain relatively constant over the period of $2 \cdot N$ chips. However, the value of N should be selected to be large enough to allow for an accurate estimate of the channel response of each multipath to be processed.

**[0095]** FIG. 7 is a flow diagram of a process 700 to derive the total pilot interference for a number of multipaths, in accordance with an embodiment of the invention. Process 700 may be implemented by the finger processor shown in FIG. 5.

**[0096]** Initially, the accumulator used to accumulate the estimated pilot interferences is cleared, at step 712. An interfering multipath that has not been processed is then selected, at step 714. Typically, the pilot interference is estimated for each multipath assigned for data demodulation. However, pilot interference due to unassigned multipaths may also be estimated. In general, any number of

interfering multipaths may be processed, and these multipaths are those for which the pilot interference is to be estimated and accumulated to derive the total pilot interference.

**[0097]** The data samples for the received signal with the selected multipath is then processed to derive an estimate of the channel response of the selected multipath, at step 716. The channel response may be estimated based on the pilot in the selected multipath, as described above. For cdma2000, this processing entails (1) spreading the data samples with a spreading sequence for the multipath (i.e., with the proper phase corresponding to the time offset of the multipath), (2) channelizing the despread data samples to provide pilot symbols (e.g., multiplying the despread samples with the pilot channelization code and accumulating the channelized data samples over the pilot channelization code length), and (3) filtering the pilot symbols to derive pilot estimates that are indicative of the channel response of the selected multipath. Estimation of the channel response based on some other techniques may also be used, and this is within the scope of the invention.

**[0098]** The pilot interference due to the selected multipath is then estimated, at step 718. The pilot interference may be estimated by generating processed pilot data and multiplying this data with the estimated channel response derived in step 716. The processed pilot data is simply the spreading sequence for the selected multipath if the pilot data is a sequence of all zeros and the pilot channelization code is also all zeros. In general, the processed pilot data is the data after all signal processing at the transmitter unit but prior to the filtering and frequency upconversion (e.g., the data at the output of modulator 216a in FIG. 3 for the reverse link in cdma2000).

**[0099]** The estimated pilot interference for the selected multipath is then accumulated in the interference accumulator with the estimated pilot interferences for prior-processed multipaths, at step 720. As noted above, the timing phase of the multipath is observed in performing steps 716, 718, and 720.

**[00100]** A determination is then made whether or not all interfering multipaths have been processed, at step 722. If the answer is no, then the process returns to step 714 and another interfering multipath is selected for processing. Otherwise, the

content of the accumulator represents the total pilot interference due to all processed multipath, which may be provided in step 724. The process then terminates.

**[00101]** The pilot interference estimation in FIG. 7 may be performed for all multipaths in a time-division multiplexed manner using one or more finger processors. Alternatively, the pilot interference estimation for multiple multipaths may be performed in parallel using a number of finger processors. In this case, if the hardware has sufficient capabilities, then the pilot interference estimation and cancellation may be performed in real-time along with the data demodulation (e.g., as the data samples are received, with minimal or no buffering, as described above).

**[00102]** FIG. 8 is a flow diagram of a process 800 to data demodulate a number of multipaths with pilot interference cancellation, in accordance with an embodiment of the invention. Process 800 may also be implemented by the finger processor shown in FIG. 5.

**[00103]** Initially, the total pilot interference due to all multipaths of interest is derived, at step 812. Step 812 may be implemented using process 700 shown in FIG. 7. A particular multipath is then selected for data demodulation, at step 814. In an embodiment and as described above, the total pilot interference is initially canceled from the selected multipath, at step 816. This may be achieved by subtracting the interference samples for the total pilot interference (which are stored in the accumulator) from the data samples for the received signal that includes the selected multipath.

**[00104]** Data demodulation is then performed on the pilot-canceled signal in the normal manner. For cdma2000, this entails (1) despreading the pilot-canceled data samples, (2) channelizing the despread data to provide data symbols, and (3) demodulating the data symbols with the pilot estimates. The demodulated symbols (i.e., the demodulated data) for the selected multipath are then combined with the demodulated symbols for other multipaths for the same transmitter unit (e.g., terminal). The demodulated symbols for multipaths in multiple received signals (e.g., if receive diversity is employed) may also be combined. The symbol combining may be achieved by the symbol combiner shown in FIG. 4.

**[00105]** A determination is then made whether or not all assigned multipaths have been demodulated, at step 822. If the answer is no, then the process returns to step 814 and another multipath is selected for data demodulation. Otherwise, the process terminates.

**[00106]** As noted above, the data demodulation for all assigned multipaths of a given transmitter unit may be performed in a time-division multiplexed manner using one or more finger processors. Alternatively, the data demodulation for all assigned multipaths may be performed in parallel using a number of finger processors.

**[00107]** Referring back to FIGS. 4 and 5, searcher 412 may be designed and operated to search for new multipaths based on the pilot-canceled data samples (instead of the raw received data samples from buffers 408). This may provided improved search performance since the pilot interference from some or all known multipaths may have been removed as described above.

**[00108]** The pilot interference cancellation techniques described herein may be able to provide noticeable improvement in performance. The pilot transmitted by each terminal on the reverse link contributes to the total channel interference, $I_o$, in similar manner as background noise, $N_o$. The pilots from all terminals may represent a substantial part of the total interference level seen by all terminals. This would then result in a lower signal-to-total-noise-plus-interference ratio (SNR) for the individual terminal. In fact, it is estimated that in a cdma2000 system (which supports pilots on the reverse link) operating near capacity, approximately half of the interference seen at a base station may be due to the pilots from the transmitting terminals. Cancellation of the pilot interference may thus improve the SNR of each individual terminal, which then allows each terminal to transmit at a lower power level and increase the reverse link capacity.

**[00109]** The techniques described herein for estimating and canceling pilot interference may be advantageously used in various wireless communication systems that transmit a pilot along with data. For example, these techniques may be used for various CDMA systems (e.g., cdma2000, IS-95, W-CDMA, TS-CDMA, and so on), Personal Communication Services (PCS) systems (e.g., ANSI J-STD-008), and other wireless communication systems. The techniques described herein may be used to

estimate and cancel pilot interference in cases where multiple instances of each of one or more transmitted signals are received and processed (e.g., by a rake receiver or some other demodulator) and also in cases where multiple transmitted signals are received and processed.

**[00110]** For clarity, various aspects and embodiments of the invention have been described for the reverse link in cdma2000. The pilot interference cancellation techniques described herein may also be used for the forward link from the base station to the terminal. The processing by the demodulator is determined by the particular CDMA standard being supported and whether the inventive techniques are used for the forward or reverse link. For example, the "despreading" with a spreading sequence in IS-95 and cdma2000 is equivalent to the "descrambling" with a scrambling sequence in W-CDMA, and the channelization with a Walsh code or a quasi-orthogonal function (QOF) in IS-95 and cdma2000 is equivalent to the "despreading" with an OVSF code in W-CDMA. In general, the processing performed by the demodulator at the receiver is complementary to that performed by the modulator at the transmitter unit.

**[00111]** For the forward link, the techniques described herein may also be used to approximately cancel other pilots that may be transmitted in addition to, or possibly in place of, a "common" pilot transmitted to all terminals in a cell. For example, cdma2000 supports a "transmit diversity" pilot and an "auxiliary" pilot. These other pilots may utilize different Walsh codes (i.e., different channelization codes, which may be quasi-orthogonal functions). A different data pattern may also be used for the pilot. To process any of these pilots, the despread samples are decovered with the same Walsh code used to channelize the pilot at the base station, and further correlated (i.e., multiplied and accumulated) with the same pilot data pattern used at the base station for the pilot. The transmit diversity pilot and/or auxiliary pilot may be estimated and canceled in addition to the common pilot.

**[00112]** Similarly, W-CDMA supports a number of different pilot channels. First, a common pilot channel (CPICH) may be transmitted on a primary base station antenna. Second, a diversity CPICH may be generated based on non-zero pilot data and transmitted on a diversity antenna of the base station. Third, one or more

secondary CPICHs may be transmitted in a restricted part of the cell, and each secondary CPICH is generated using a non-zero channelization code. Fourth, the base station may further transmit a dedicated pilot to a specific user using the same channelization code as the user's data channel. In this case, the pilot symbols are time-multiplexed with the data symbols to that user. Accordingly, it will be understood by those skilled in the art that the techniques described herein are applicable for processing all of the above different types of pilot channels, and other pilot channels that may also be transmitted in a wireless communication system.

**[00113]** The demodulator and other processing units that may be used to implement various aspects and embodiments of the invention may be implemented in hardware, software, firmware, or a combination thereof. For a hardware design, the demodulator (including the data demodulation unit and the elements used for pilot interference estimation and cancellation such as the pilot estimator and the pilot interference estimator), and other processing units may be implemented within one or more application specific integrated circuits (ASIC), digital signal processors (DSP), digital signal processing devices (DSPDs), field programmable gate arrays (FPGA), processors, microprocessors, controllers, microcontrollers, programmable logic devices (PLD), other electronic units, or any combination thereof.

**[00114]** For a software implementation, the elements used for pilot interference estimation and cancellation and data demodulation may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memory 262 in FIG. 2) and executed by a processor (e.g., controller 260). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as it known in the art.

**[00115]** The elements used to implement the pilot interference estimation and cancellation described herein may be incorporated in a receiver unit or a demodulator that may further be incorporated in a terminal (e.g., a handset, a handheld unit, a stand-alone unit, and so on), a base station, or some other communication devices or units. The receiver unit or demodulator may be implemented with one or more integrated circuits.

**[00116]** The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WHAT IS CLAIMED IS:

1. A method for canceling pilot interference at a receiver unit in a wireless communication system, comprising:

receiving a signal comprised of a plurality of signal instances, wherein each signal instance includes a pilot;

deriving total pilot interference due to one or more signal instances;

subtracting the total pilot interference from the received signal to derive a pilot-canceled signal; and

processing the pilot-canceled signal to derive demodulated data for each of at least one signal instance in the received signal.

2. The method of claim 1, wherein the total pilot interference is derived by

estimating pilot interference due to each of the one or more signal instances, and

accumulating the estimated pilot interference for the one or more signal instances.

3. The method of claim 2, wherein the pilot interference due to each of the one or more signal instances is estimated by

processing the signal instance to derive an estimate of a channel response of the signal instance, and

multiplying processed pilot data for the signal instance with the estimated channel response to provide the estimated pilot interference.

4. The method of claim 3, wherein the processed pilot data for each of the one or more signal instances is a spreading sequence for the signal instance.

5. The method of claim 4, wherein the spreading sequence for the signal instance has a phase corresponding to an arrival time of the signal instance.

6. The method of claim 3, wherein the estimated channel response for each of the one or more signal instances is derived by

despreading data samples for the received signal with a spreading sequence for the signal instance,

channelizing the despread samples with a pilot channelization code to provide pilot symbols, and

filtering the pilot symbols to provide the estimated channel response.

7. The method of claim 3, wherein the estimated channel response of the signal instance is derived based on a current segment of data samples for the received signal and the estimated pilot interference is for a subsequent segment of data samples.

8. The method of claim 3, wherein the estimated channel response of the signal instance is derived based on a current segment of data samples for the received signal and the estimated pilot interference is for the same segment of data samples.

9. The method of claim 3, wherein the estimated channel response for each of the one or more signal instances is derived based on data samples for the received signal.

10. The method of claim 3, wherein the estimated channel response for each of the one or more signal instances is derived based on data samples having pilot from the signal instance unremoved but pilots from other interfering signal instances removed.

11. The method of claim 1, wherein the processing of the pilot-canceled signal for each of the at least one signal instance includes

despreading samples for the pilot-canceled signal with a spreading sequence for the signal instance,

channelizing the despread samples with a data channelization code to provide data symbols, and

demodulating the data symbols with pilot estimates to provide the demodulated data for the signal instance.

12. The method of claim 11, wherein the pilot estimates for each of the at least one signal instance are derived based on data samples for the received signal.

13. The method of claim 11, wherein the pilot estimates for each of the at least one signal instance are derived based on data samples having pilot from the signal instance unremoved but pilots from other interfering signal instances removed.

14. The method of claim 2, wherein the pilot interference due to the one or more signal instances is estimated in a time-division multiplexed manner.

15. The method of claim 1, wherein the subtracting includes

subtracting interference samples for the total pilot interference from data samples for the received signal, wherein the interference samples and data samples are both provided at a particular sample rate.

16. The method of claim 1, wherein the pilot interference due to a signal instance being processed to derive the demodulated data is excluded from the total pilot interference.

17. The method of claim 1, further comprising:

processing the pilot-canceled signal to search for new signal instances in the received signal.

18. The method of claim 15, wherein the sample rate is multiple times a chip rate.

19. The method of claim 1, wherein the deriving the total pilot interference is performed based on segments of data samples for the received signal.

20. The method of claim 19, wherein the each segment includes data samples for one symbol period.

21. The method of claim 1, wherein the processing to derive demodulated data is performed based on segments of pilot-canceled data samples for the pilot-canceled signal.

22. The method of claim 1, wherein the deriving the total pilot interference and the processing of the pilot-canceled signal are performed in parallel.

23. The method of claim 1, wherein the deriving the total pilot interference and the processing of the pilot-canceled signal are performed in a pipelined manner.

24. The method of claim 1, wherein the wireless communication system is a CDMA system.

25. The method of claim 24, wherein the CDMA system supports cdma2000 standard.

26. The method of claim 24, wherein the CDMA system supports W-CDMA standard.

27. The method of claim 24, wherein the CDMA system supports IS-95 standard.

28. The method of claim 24, wherein the received signal comprises one or more reverse link modulated signals in the CDMA system.

29. The method of claim 24, wherein the received signal comprises one or more forward link modulated signals in the CDMA system.

30. A method for canceling pilot interference at a receiver unit in a wireless communication system, comprising:

processing a received signal comprised of a plurality of signal instances to provide data samples, wherein each signal instance includes a pilot;

processing the data samples to derive an estimate of pilot interference due to each of one or more signal instances;

deriving total pilot interference due to the one or more signal instances based on the estimated pilot interference;

subtracting the total pilot interference from the data samples to derive pilot-canceled data samples; and

processing the pilot-canceled data samples to derive demodulated data for each of at least one signal instance in the received signal.

31. The method of claim 30, wherein the processing the data samples to derive the estimated pilot interference due to each of the one or more signal instances includes

despreading the data samples with a spreading sequence for the signal instance,

channelizing the despread samples with a pilot channelization code to provide pilot symbols,

filtering the pilot symbols to provide an estimate or a channel response of the signal instance, and

multiplying the spreading sequence for the signal instance with the estimated channel response to provide the estimated pilot interference due to the signal instance.

32. The method of claim 30, wherein the processing the pilot-canceled data samples to derive the demodulated data for each of the at least one signal instance includes

despreading the pilot-canceled data samples with a spreading sequence for the signal instance,

channelizing the despread samples with a data channelization code to provide data symbols, and

demodulating the data symbols to provide the demodulated data for the signal instance.

33. The method of claim 30, wherein the subtracting includes

subtracting interference samples for the total pilot interference from the data samples for the received signal, wherein the interference samples

and data samples are both provided at a particular sample rate that is multiple times a chip rate.

34. A receiver unit in a wireless communication system, comprising:

a receiver configured to process a received signal comprised of a plurality of signal instances to provide data samples, wherein each signal instance includes a pilot; and

a demodulator including

a pilot interference estimator configured to process the data samples to derive an estimate of pilot interference due to each of one or more signal instances and to derive total pilot interference due to the one or more signal instances based on the estimated pilot interference,

a summer configured to subtract the total pilot interference from the data samples to derive pilot-canceled data samples, and

a data demodulation unit configured to process the pilot-canceled data samples to derive demodulated data for each of at least one signal instance in the received signal.

35. The receiver unit of claim 34, wherein the demodulator further includes

a channel estimator configured to provide an estimated channel response for each of the one or more signal instances.

36. The receiver unit of claim 35, wherein the pilot interference estimator is further configured to multiply processed pilot data for each of the one or more signal instances with the estimated channel response for the signal instance to provide the estimated pilot interference due to the signal instance.

37. The receiver unit of claim 34, wherein for each of the at least one signal instance the data demodulation unit is configured to despread the pilot-canceled data samples with a spreading sequence for the signal instance, channelize the despread samples with a data channelization code to provide data symbols, and demodulate the data symbols with pilot estimates for the signal instance to provide the demodulated data for the signal instance.

38. A terminal in a CDMA system comprising:

a receiver configured to process a received signal comprised of a plurality of signal instances to provide data samples, wherein each signal instance includes a pilot; and

a demodulator including

a pilot interference estimator configured to process the data samples to derive an estimate of pilot interference due to each of one or more signal instances and to derive total pilot interference due to the one or more signal instances based on the estimated pilot interference,

a summer configured to subtract the total pilot interference from the data samples to derive pilot-canceled data samples, and

a data demodulation unit configured to process the pilot-canceled data samples to derive demodulated data for each of at least one signal instance in the received signal.

39. The terminal of claim 38, wherein the demodulator further includes

a channel estimator configured to provide an estimated channel response for each of the one or more signal instances.

40. The terminal of claim 39, wherein the pilot interference estimator is further configured to multiply processed pilot data for each of the one or

more signal instances with the estimated channel response for the signal instance to provide the estimated pilot interference due to the signal instance.

41. The terminal of claim 38, wherein for each of the at least one signal instance the data demodulation unit is configured to despread the pilot-canceled data samples with a spreading sequence for the signal instance, channelize the despread samples with a data channelization code to provide data symbols, and demodulate the data symbols with pilot estimates for the signal instance to provide the demodulated data for the signal instance.

42. A base station in a CDMA system comprising:

a receiver configured to process a received signal comprised of a plurality of signal instances to provide data samples, wherein each signal instance includes a pilot; and

a demodulator including

a pilot interference estimator configured to process the data samples to derive an estimate of pilot interference due to each of one or more signal instances and to derive total pilot interference due to the one or more signal instances based on the estimated pilot interference,

a summer configured to subtract the total pilot interference from the data samples to derive pilot-canceled data samples, and

a data demodulation unit configured to process the pilot-canceled data samples to derive demodulated data for each of at least one signal instance in the received signal.

43. The base station of claim 42, wherein the demodulator further includes

a channel estimator configured to provide an estimated channel response for each of the one or more signal instances.

44. The base station of claim 43, wherein the pilot interference estimator is further configured to multiply processed pilot data for each of the one or more signal instances with the estimated channel response for the signal instance to provide the estimated pilot interference due to the signal instance.

45. The base station of claim 42, wherein for each of the at least one signal instance the data demodulation unit is configured to despread the pilot-canceled data samples with a spreading sequence for the signal instance, channelize the despread samples with a data channelization code to provide data symbols, and demodulate the data symbols with pilot estimates for the signal instance to provide the demodulated data for the signal instance.

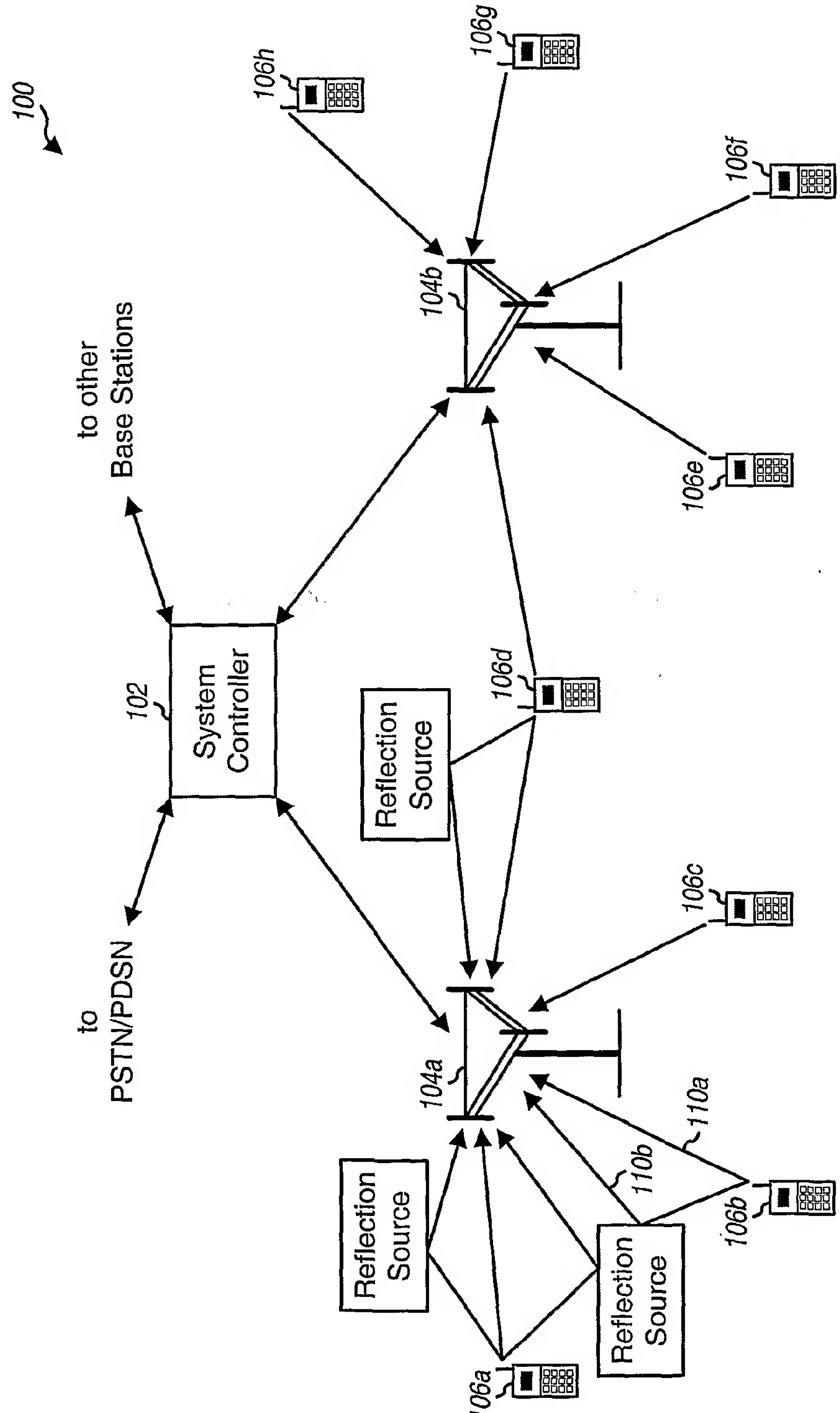

100
to other
Base Stations
to
PSTN/PDSN
102
System
Controller
104b
106h
106g
106f
106e
106d
Reflection
Source
106c
104a
110a
110b
106b
Reflection
Source
Reflection
Source
106a

FIG. 1

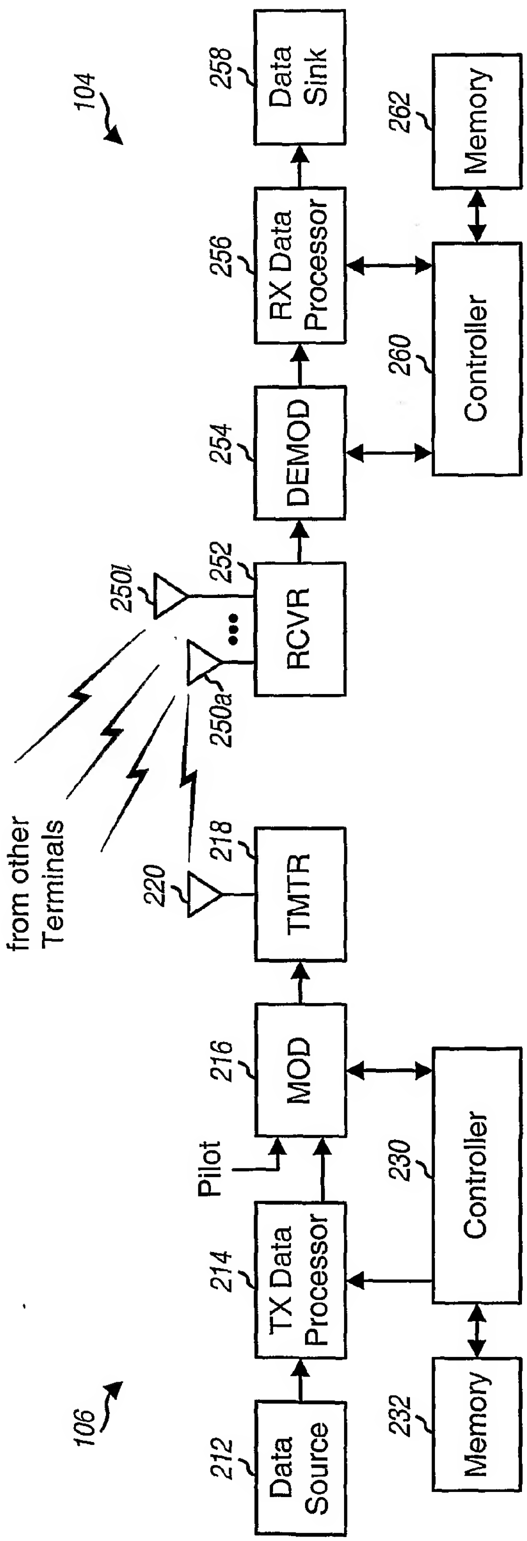

104
106
Data Source
212
TX Data Processor
214
Pilot
MOD
216
TMTR
218
220
from other Terminals
250a
250l
RCVR
252
DEMOD
254
RX Data Processor
256
Data Sink
258
Controller
230
Memory
232
Controller
260
Memory
262

FIG. 2

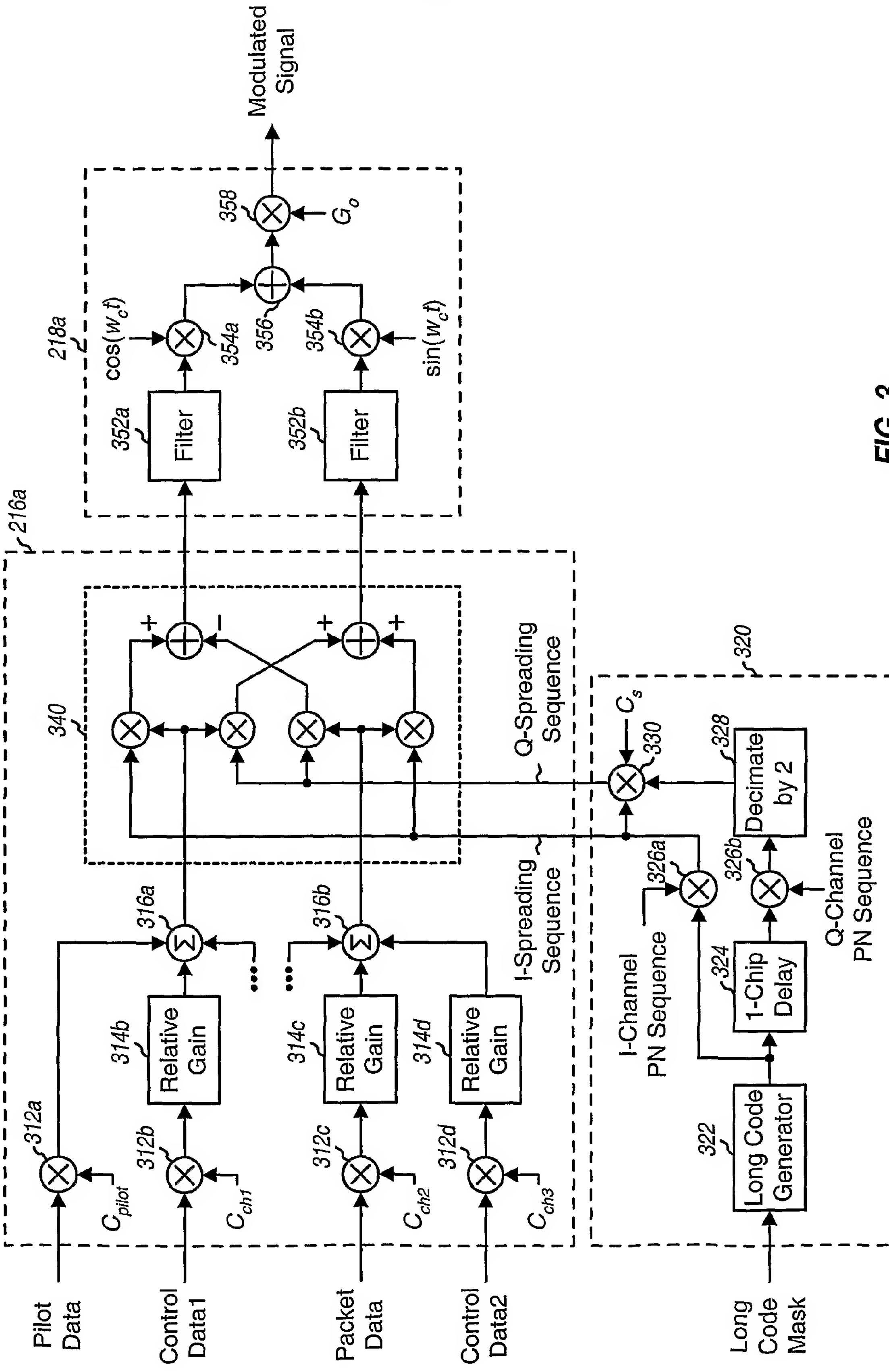

Modulated Signal
218a
cos($w_c t$)
354a
356
354b
358
$G_o$
sin($w_c t$)
352a
Filter
352b
Filter
216a
340
Q-Spreading Sequence
I-Spreading Sequence
320
$C_s$
330
328
Decimate by 2
326a
326b
I-Channel PN Sequence
Q-Channel PN Sequence
324
1-Chip Delay
322
Long Code Generator
316a
316b
314b
Relative Gain
314c
Relative Gain
314d
Relative Gain
312a
312b
312c
312d
$C_{pilot}$
$C_{ch1}$
$C_{ch2}$
$C_{ch3}$
Pilot Data
Control Data1
Packet Data
Control Data2
Long Code Mask

FIG. 3

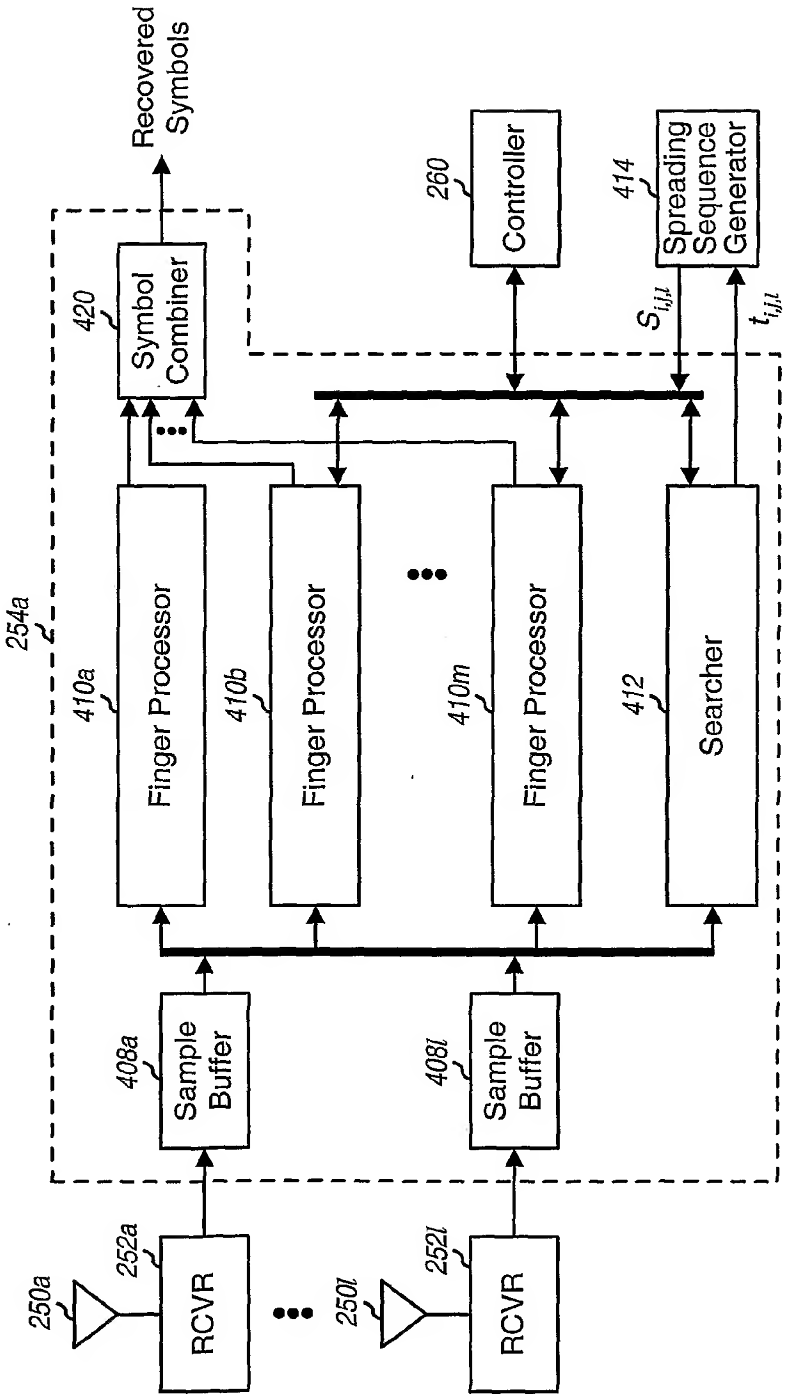

Recovered Symbols
420
Symbol Combiner
260
Controller
414
Spreading Sequence Generator
$S_{i,j,l}$
$t_{i,j,l}$
254a
410a
Finger Processor
410b
Finger Processor
410m
Finger Processor
412
Searcher
408a
Sample Buffer
408l
Sample Buffer
250a
252a
RCVR
250l
252l
RCVR

FIG. 4

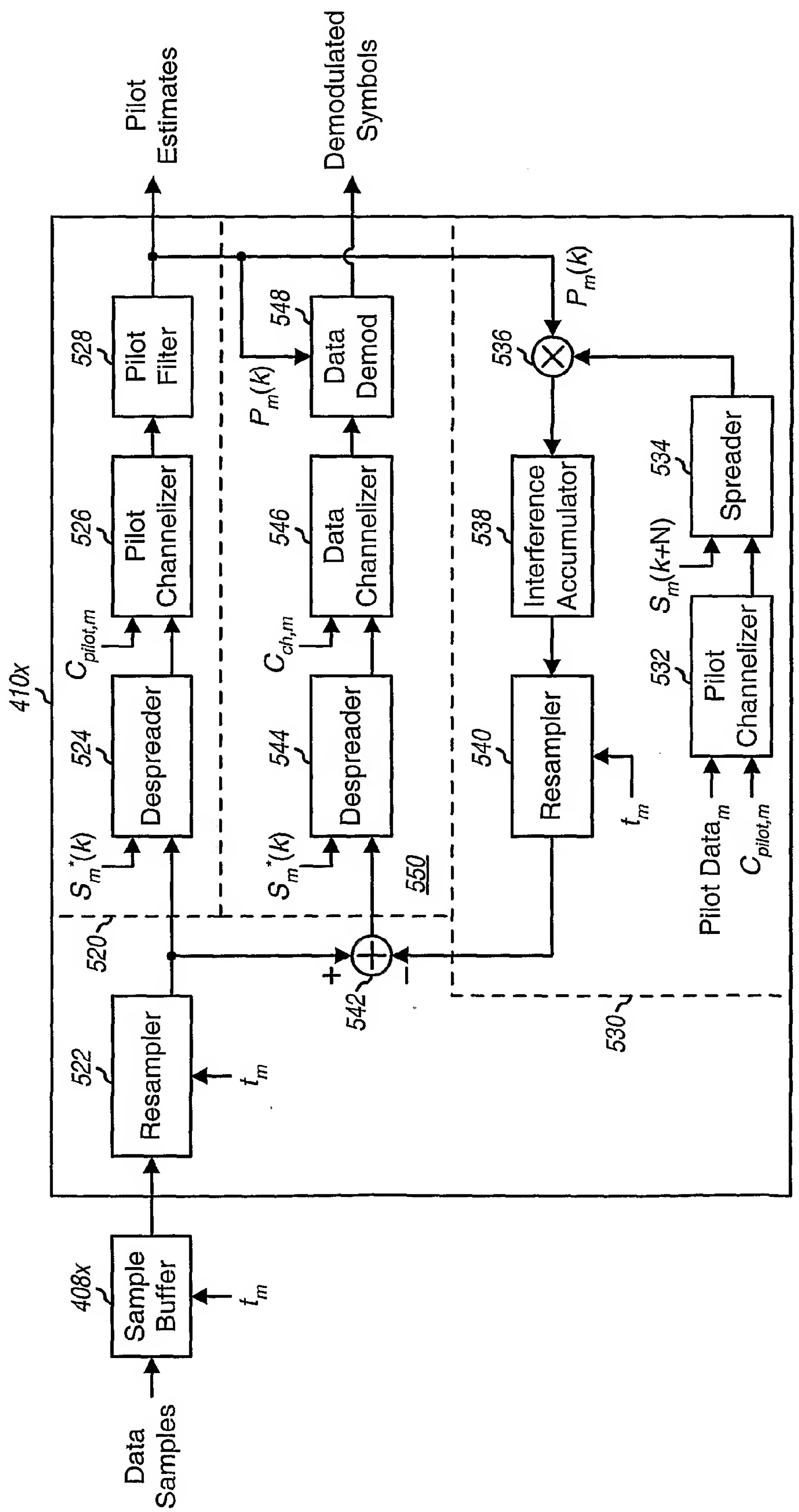
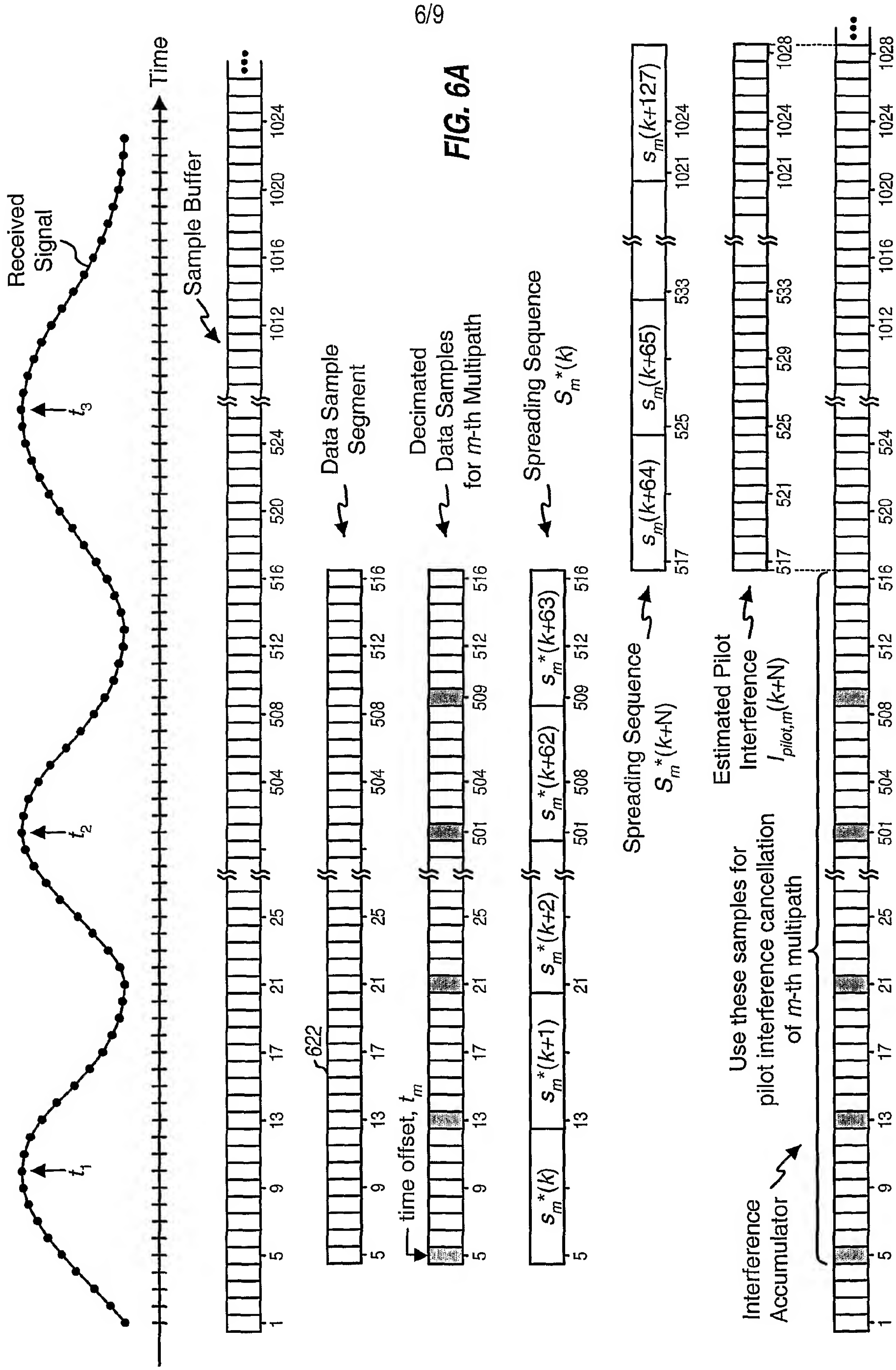

Data Samples
Sample Buffer
408x
$t_m$
410x
Resampler
522
520
Despreader
524
$S_m^*(k)$
$C_{pilot,m}$
Pilot Channelizer
526
Pilot Filter
528
Pilot Estimates
Despreader
544
$S_m^*(k)$
$C_{ch,m}$
Data Channelizer
546
Data Demod
548
$P_m(k)$
Demodulated Symbols
550
+
−
542
Resampler
540
Interference Accumulator
538
536
$P_m(k)$
Spreader
534
$S_m(k+N)$
Pilot Channelizer
532
Pilot $Data_m$
$C_{pilot,m}$
530

FIG. 5



Received Signal
Time
$t_1$
$t_2$
$t_3$
Sample Buffer
Data Sample Segment
622
time offset, $t_m$
Decimated Data Samples for m-th Multipath
Spreading Sequence $S_m^*(k)$
$s_m^*(k)$
$s_m^*(k+1)$
$s_m^*(k+2)$
$s_m^*(k+62)$
$s_m^*(k+63)$
Spreading Sequence $S_m^*(k+N)$
$s_m(k+64)$
$s_m(k+65)$
$s_m(k+127)$
Estimated Pilot Interference $I_{pilot,m}(k+N)$
Interference Accumulator
Use these samples for pilot interference cancellation of m-th multipath


FIG. 6A

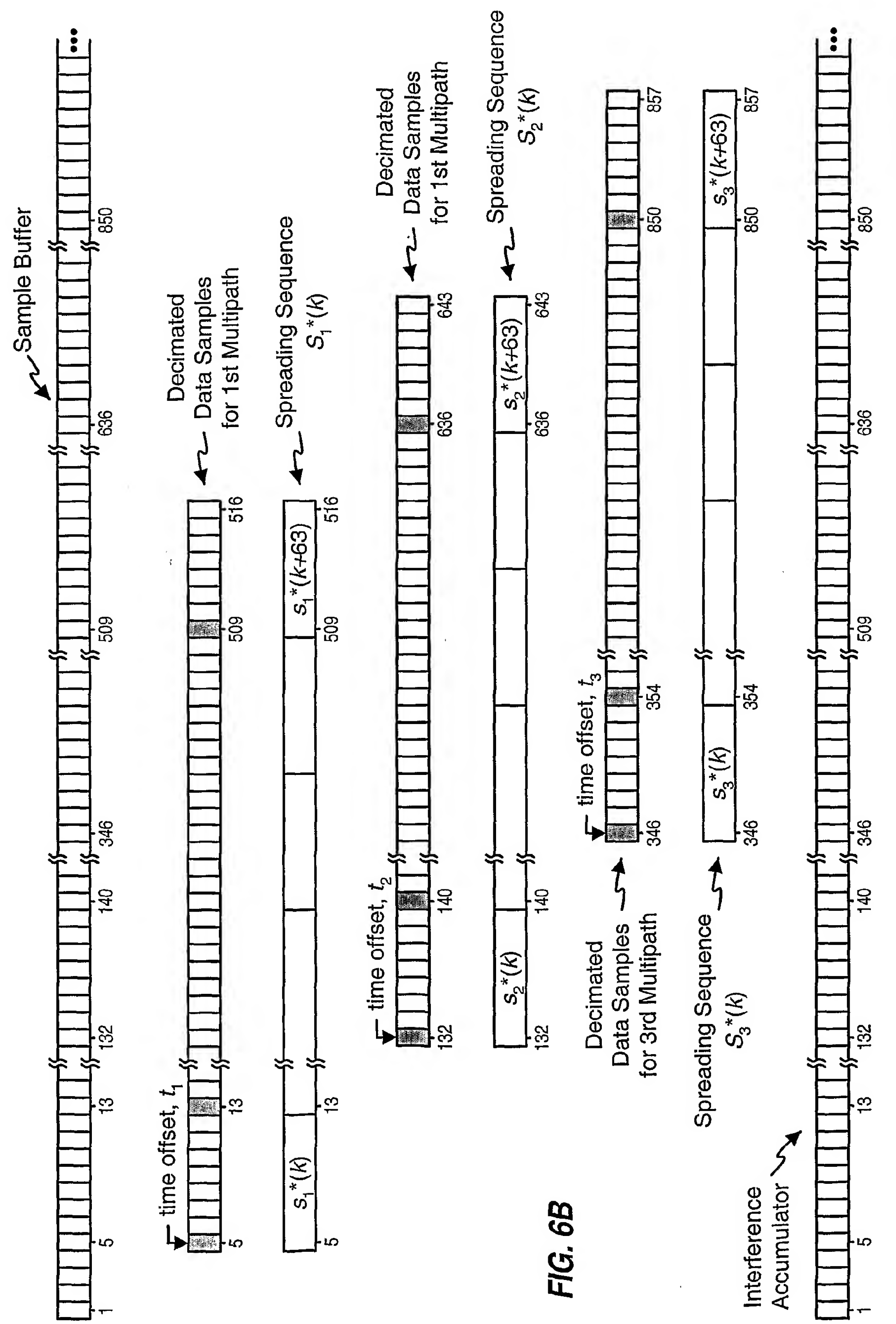
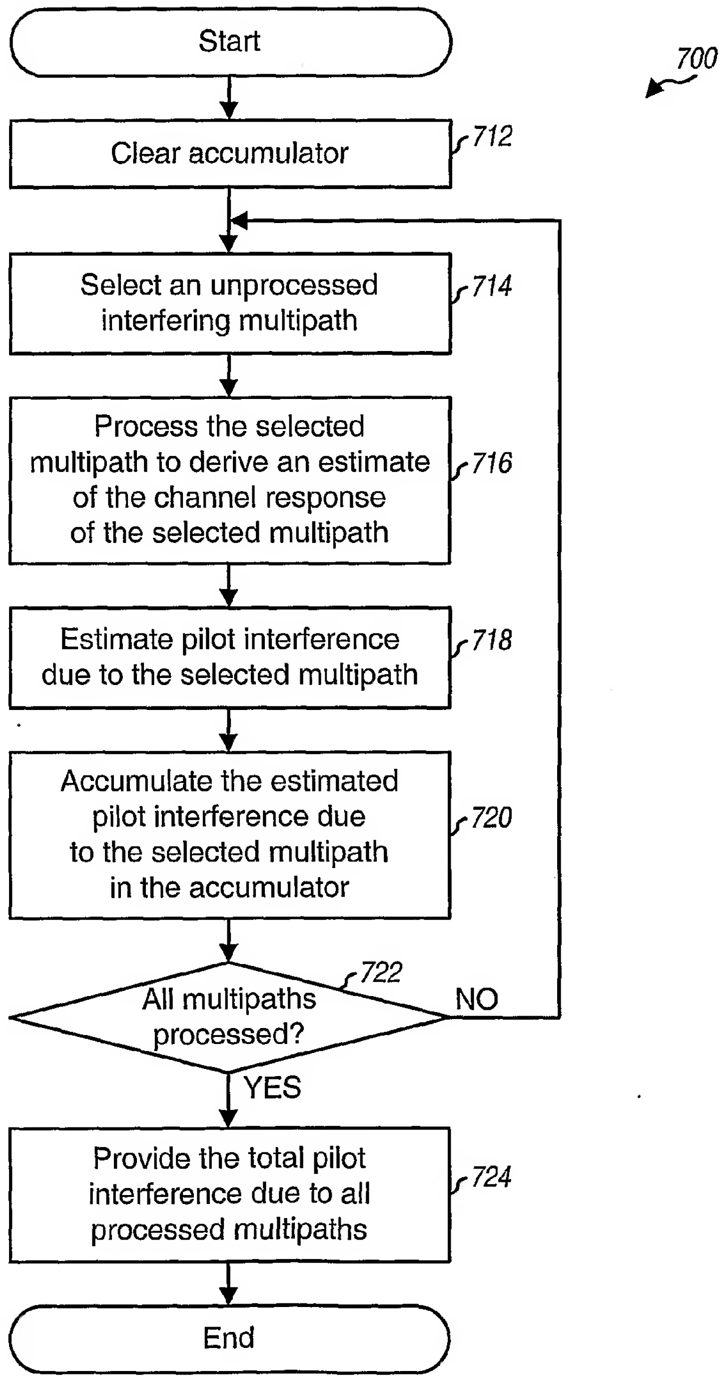

Sample Buffer
1
5
13
132
140
346
509
636
850
time offset, $t_1$
Decimated Data Samples for 1st Multipath
516
Spreading Sequence $S_1^*(k)$
$s_1^*(k)$
$s_1^*(k+63)$
time offset, $t_2$
Decimated Data Samples for 1st Multipath
643
Spreading Sequence $S_2^*(k)$
$s_2^*(k)$
$s_2^*(k+63)$
time offset, $t_3$
Decimated Data Samples for 3rd Multipath
354
857
Spreading Sequence $S_3^*(k)$
$s_3^*(k)$
$s_3^*(k+63)$
Interference Accumulator

FIG. 6B



700
Start
Clear accumulator
712
Select an unprocessed interfering multipath
714
Process the selected multipath to derive an estimate of the channel response of the selected multipath
716
Estimate pilot interference due to the selected multipath
718
Accumulate the estimated pilot interference due to the selected multipath in the accumulator
720
All multipaths processed?
722
NO
YES
Provide the total pilot interference due to all processed multipaths
724
End


FIG. 7

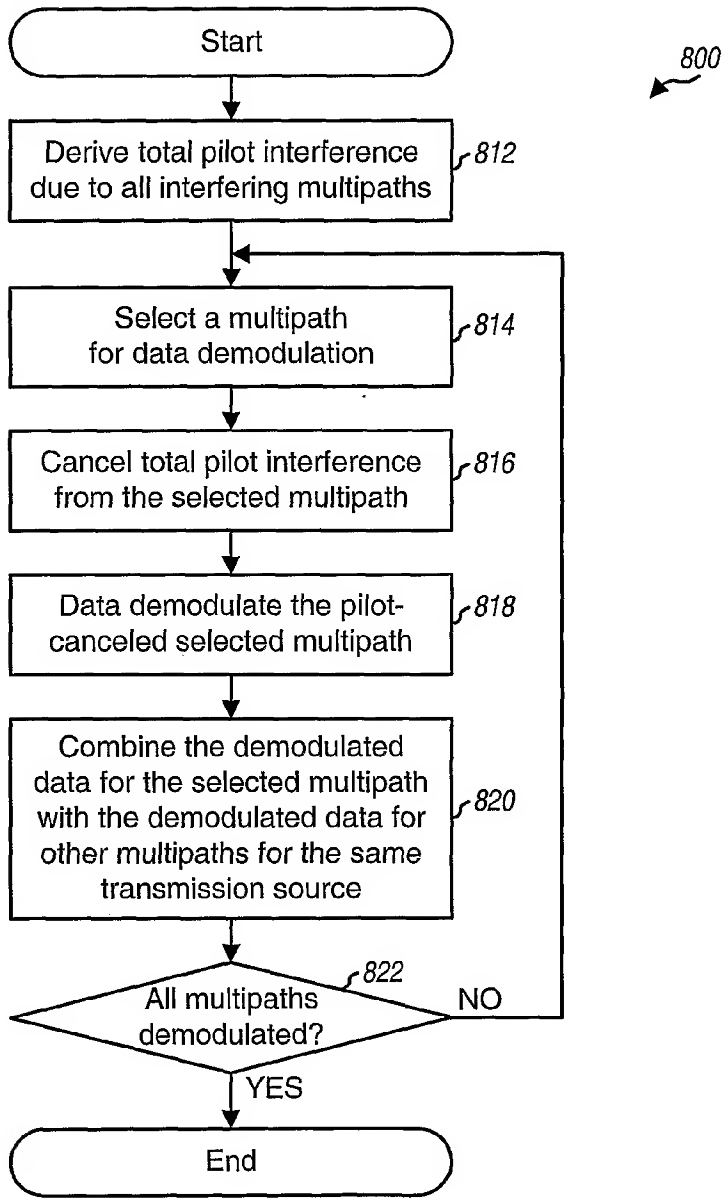

800
Start
Derive total pilot interference due to all interfering multipaths
812
Select a multipath for data demodulation
814
Cancel total pilot interference from the selected multipath
816
Data demodulate the pilot-canceled selected multipath
818
Combine the demodulated data for the selected multipath with the demodulated data for other multipaths for the same transmission source
820
All multipaths demodulated?
822
NO
YES
End

FIG. 8

# INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 02/18133

**A. CLASSIFICATION OF SUBJECT MATTER**
IPC 7 H04B1/707

According to International Patent Classification (IPC) or to both national classification and IPC

**B. FIELDS SEARCHED**

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

INSPEC, EPO-Internal

**C. DOCUMENTS CONSIDERED TO BE RELEVANT**

| Category ° | Citation of document, with indication, where appropriate, of the relevant passages | Relevant to claim No. |
|---|---|---|
| X | US 6 067 292 A (BRINK STEPHAN TEN ET AL) 23 May 2000 (2000-05-23) abstract; figures 6,7,10,12,14,16,16S,17,22,23 column 2, line 1 - line 42 column 7, line 4 -column 12, line 28 column 15, line 10 - line 22 | 1-5 |

-/--

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

| Date of the actual completion of the international search | Date of mailing of the international search report |
|---|---|
| 19 September 2002 | 30/09/2002 |

| Name and mailing address of the ISA | Authorized officer |
|---|---|
| European Patent Office, P.B. 5818 Patentlaan 2<br>NL – 2280 HV Rijswijk<br>Tel. (+31–70) 340–2040, Tx. 31 651 epo nl,<br>Fax: (+31–70) 340–3016 | Bauer, F |

# INTERNATIONAL SEARCH REPORT

Int al Application No
PCT/US 02/18133

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

| Category ° | Citation of document, with indication,where appropriate, of the relevant passages | Relevant to claim No. |
|---|---|---|
| X | IWAKIRI N: "INTERFERENCE REDUCTION EFFICIENCY OF A TURBO CODED CDMA MULTILAYER SYSTEM EQUIPPED WITH A PILOT CANCELER" VTC 1999-FALL. IEEE VTS 50TH. VEHICULAR TECHNOLOGY CONFERENCE. GATEWAY TO THE 21ST. CENTURY COMMUNICATIONS VILLAGE. AMSTERDAM, SEPT. 19 - 22, 1999, IEEE VEHICULAR TECHNOLGY CONFERENCE, NEW YORK, NY: IEEE, US, vol. 1 CONF. 50, September 1999 (1999-09), pages 391-395, XP000929078 ISBN: 0-7803-5436-2 paragraph '000C!; figure 3 | 1 |
| X | EP 0 980 149 A (IND TECH RES INST) 16 February 2000 (2000-02-16) paragraphs '0011!,'0017!,'0026!-'0032!; figure 2A | 1-5 |

# INTERNATIONAL SEARCH REPORT

tional application No.
PCT/US 02/18133

**Box I Observations where certain claims were found unsearchable (Continuation of item 1 of first sheet)**

This International Search Report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. [ ] Claims Nos.:
because they relate to subject matter not required to be searched by this Authority, namely:

2. [X] Claims Nos.: 6-45
because they relate to parts of the International Application that do not comply with the prescribed requirements to such an extent that no meaningful International Search can be carried out, specifically:
see FURTHER INFORMATION sheet PCT/ISA/210

3. [ ] Claims Nos.:
because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a).

**Box II Observations where unity of invention is lacking (Continuation of item 2 of first sheet)**

This International Searching Authority found multiple inventions in this international application, as follows:

1. [ ] As all required additional search fees were timely paid by the applicant, this International Search Report covers all searchable claims.

2. [ ] As all searchable claims could be searched without effort justifying an additional fee, this Authority did not invite payment of any additional fee.

3. [ ] As only some of the required additional search fees were timely paid by the applicant, this International Search Report covers only those claims for which fees were paid, specifically claims Nos.:

4. [ ] No required additional search fees were timely paid by the applicant. Consequently, this International Search Report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

**Remark on Protest**

[ ] The additional search fees were accompanied by the applicant's protest.

[ ] No protest accompanied the payment of additional search fees.

FURTHER INFORMATION CONTINUED FROM PCT/ISA/ 210

Continuation of Box I.2

Claims Nos.: 6-45

In view of the large number of independent claims (5), and on the even larger number of claims dependent on the not novel calims 1 and 3 (16) , which render it difficult, if not impossible, to determine the matter for which protection is sought, the present application fails to comply with the clarity and conciseness requirements of Article 6 PCT (see also Rule 6.1(a) PCT) to such an extent that a meaningful search is impossible. Consequently, the search has been carried out for those parts of the application which do appear to be clear (and concise), namely claims 1-5.

The applicant's attention is drawn to the fact that claims, or parts of claims, relating to inventions in respect of which no international search report has been established need not be the subject of an international preliminary examination (Rule 66.1(e) PCT). The applicant is advised that the EPO policy when acting as an International Preliminary Examining Authority is normally not to carry out a preliminary examination on matter which has not been searched. This is the case irrespective of whether or not the claims are amended following receipt of the search report or during any Chapter II procedure.

# INTERNATIONAL SEARCH REPORT

Information on patent family members

Int  l Application No
PCT/US 02/18133

| Patent document cited in search report | | Publication date | Patent family member(s) | | Publication date |
|---|---|---|---|---|---|
| US 6067292 | A | 23-05-2000 | US | 6009089 A | 28-12-1999 |
| | | | EP | 0876002 A2 | 04-11-1998 |
| | | | JP | 10327126 A | 08-12-1998 |
| | | | KR | 263801 B1 | 16-08-2000 |
| EP 0980149 | A | 16-02-2000 | US | 6154443 A | 28-11-2000 |
| | | | EP | 0980149 A2 | 16-02-2000 |
| | | | TW | 419912 B | 21-01-2001 |

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
6 February 2003 (06.02.2003)

PCT

(10) International Publication Number
WO 03/010984 A1

(51) International Patent Classification[7]: H04Q 7/22, H04L 12/28, 12/56, H04B 7/26, H04Q 7/38

(21) International Application Number: PCT/CA02/00941

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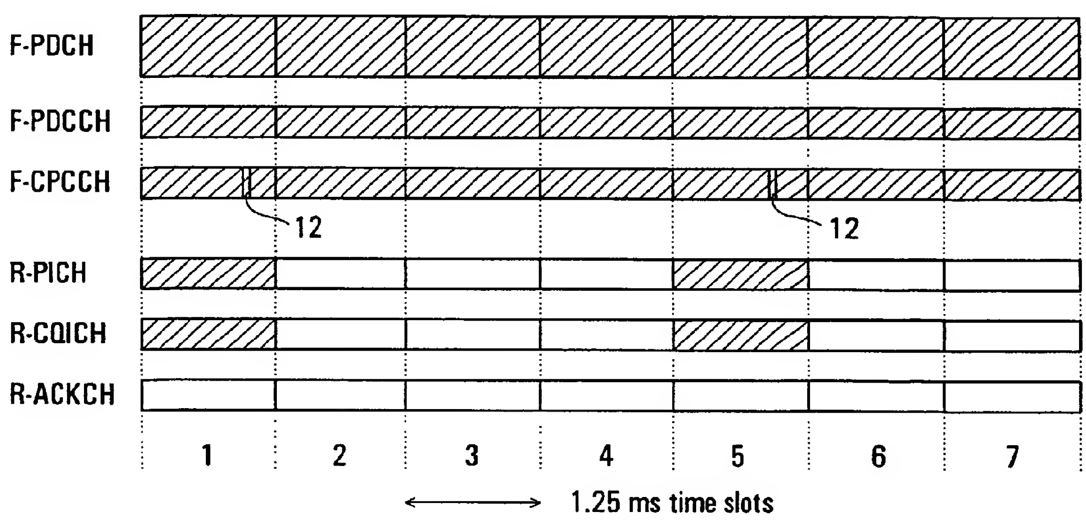
(54) **Title:** COMMUNICATION OF CONTROL INFORMATION IN WIRELESS COMMUNICATION SYSTEMS




(57) **Abstract:** A wireless system has a high rate data channel (F-PDCH) for time multiplexed communications to multiple mobile stations (MSs). Control channels include a forward link common power control channel (F-CPCCH) and reverse link feedback channels for pilot (R-PICH), forward channel quality (R-CQICH),and data acknowledgements (R-ACKCH) from each MS. An MS can have an active state for data communications, for which these control channels are used at the full (time slot) rate, or a control hold state, in which acknowledgements are not needed and the others of these control channels can be shared among a plurality of MSs in the control hold state and each using a reduced rate such as 1/2, 1/4, or 1/8 of the full rate. The arrangement can support an increased number of active MSs, factilitating an increased total throughput on the high rate data channel, without increasing system resources for the control channels.

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# COMMUNICATION OF CONTROL INFORMATION IN WIRELESS COMMUNICATION SYSTEMS

This invention relates to the communication of control information in wireless communications systems, and is particularly concerned with communication of control information (including feedback, pilot, and any other overhead information) related to supporting high rate packet data communications in a wireless communications system.

## Background

It is known in wireless communications systems to provide various channels for communications between a base station (BS) and remote stations any one of which may be mobile and accordingly is referred to as a mobile station (MS). Such channels include, for example, both dedicated and shared traffic channels and control channels, for either or both of a forward link from a BS to an MS and a reverse link from an MS to a BS.

In known wireless systems an MS can have individually assigned to it, at call set-up or subsequently, forward and/or reverse dedicated channels for packet data communications, as well as dedicated traffic and dedicated control channels. In order to save battery power, it is known to provide a control hold mode for an active user for which a data buffer has been empty for a period of time. In the control hold mode, the high data rate supplemental channel is released and the forward and reverse dedicated control channels communicate only the pilot and power control signals (i.e. data acknowledgements are not sent) at either a full rate (the time slot rate) or a reduced rate.

In more recently proposed wireless systems there are shared traffic channels, shared control channels, and common power control channels; these known systems do not provide a control hold mode.

In these more recent systems, for example high rate packet data communications are provided on a high rate forward channel, referred to as the forward packet data channel or F-PDCH, which is shared by time division multiplexing among multiple active users and which may for example have a constant RF power. For example, with time slots of 1.25 ms, such a high rate channel may be allocated to different users (MSs) in different time slots. Each such MS is assigned a MAC (medium access control, OSI Layer 2) identifier (MAC_ID) which is transmitted in corresponding time slots on a shared control channel, referred to as a forward packet data control channel or F-PDCCH, to identify the MS for which the data on the F-PDCH is intended. A forward link scheduler provides rate control by scheduling packet data to the user who has the most favourable forward link channel condition.

In a full-queue situation in which network-side data buffers for all active users are always occupied, such an arrangement can provide significant multi-user diversity gain. However, in a more realistic non-full-queue situation in which active users' buffers are not always occupied by data, the multi-user diversity gain decreases significantly because an active user's buffer may be empty although the user has the most favourable channel condition.

In response to a user requesting a packet data communications session, the system transitions the user to an active state through dedicated or shared resource assignments. The system resources involved include a sub-channel of a

forward link common power control channel or F-CPCCH, which the MS of an active user detects for the purpose of closed-loop reverse link power control, the F-PDCCH and F-PDCH referred to above, and reverse link dedicated channels R-PICH, R-CQICH, and R-ACKCH for control signalling required for supporting the forward link high rate data transmissions. When a user is in the active state, the MS sends a pilot on the reverse pilot channel R-PICH and feedback information including a channel condition (C/I or carrier-to-interference ratio report) indicating the quality of the forward link on the reverse channel quality indication channel R-CQICH, and an ACK/NAK (acknowledgement or negative acknowledgement) indication for an ARQ (automatic retransmission) function on the reverse acknowledgement channel R-ACKCH. This reverse link information is updated in each timeslot, i.e. at a rate of 800 Hz for a time slot duration of 1.25 ms. The sub-channel of the F-CPCCH is also detected at this time slot rate of for example 800 Hz.

Although it is conceivable to increase the number of active users thereby to increase the multi-user diversity gain in non-full-queue situations, the number of active users that can be supported is restricted by these limited system resources for overhead information for active users.

It is desirable to facilitate an increase in the number of active users without increasing the associated system resources for overhead information required for supporting active users.

## Summary of the Invention

Recognizing that a result of a typically bursty nature of packet data communications is that an active user's buffer may be empty for a significant portion of the time that the user is active, this invention arises from an appreciation

that it is not necessary to perform reverse link power control at the full rate (the time slot rate of e.g. 800 Hz), and it is not necessary for an active user to send feedback information at the full rate. Instead, such an active user can operate in a lower rate mode, i.e. such information can be communicated at a reduced rate, and for channels that are shared among active users the resulting saving in system resources can enable more active users to be supported. Of course, it is also evident in this case that the feedback information need not include the R-ACKCH acknowledgements, because no packet data is being sent to the MS from the empty buffer.

The lower or reduced rate mode is also referred to as a control hold mode, but it is substantially enhanced compared with the known control hold mode referred to above. The latter only provides battery power saving, and does not involve or permit any increase in the number of active users that can be supported because the control channels are still dedicated to the active users. In contrast, embodiments of this invention not only facilitate battery power saving but also facilitate dynamic sharing of the system resources among active users so that more active users may be supported without requiring additional system resources, potentially resulting in an increased forward link throughput for the system as a whole.

According to one aspect of this invention there is provided a method of communicating control information in a wireless communications system, comprising the steps of: in a first state of a terminal for traffic communication with the terminal, communicating control information with the terminal at a first rate; in a second state of the terminal, communicating at least some of said control information at a second rate which is less than the first rate; and sharing a

communication channel for the control information at the second rate for communicating control information among a plurality of terminals in the second state.

The control information communicated at the second rate in the second state of the terminal can comprise power control information for the terminal, and/or a pilot from the terminal, and/or a channel quality indication from the terminal. The second rate is desirably a sub-multiple, preferably 1/2, 14, or 1/8, of the first rate.

Where the control information is communicated in time slots of at least one communication channel of the system, preferably control information at the first rate is communicated with a terminal in the first state in each time slot on the communication channel, and control information at the second rate is communicated with a terminal in the second state in only one of every N time slots on the communication channel, where N is an integer greater than one. Conveniently N is a power of 2, for example N=2, 4, or 8.

The method can include the step of switching between the first and second states of the terminal in dependence upon whether or not a data buffer for traffic communication with the terminal is empty.

Another aspect of the invention provides a method of communicating control information in a wireless communications system having forward and reverse channels for communicating traffic and control information with a plurality of terminals, the forward channels including a time division multiplexed (tdm) channel for communicating traffic in respective time slots to respective terminals, and the reverse channels including channels for communicating pilot and feedback

information from the terminals, the method comprising the steps of: in a first state of a terminal for receiving traffic for the terminal in respective time slots of the tdm channel, communicating said pilot and feedback information at a first rate; in a second state of the terminal, communicating said pilot and feedback information at a second rate which is less than the first rate; and sharing the channels for communicating said pilot and feedback information at the second rate among a plurality of terminals in the second state.

Conveniently the first rate is equal to a rate of said time slots and the second rate is 1/2, 1/4, or 1/8 of the first rate.

Preferably in this method the forward channels include a power control channel for controlling, at the first rate, power on the reverse channels of a respective terminal in the first state, the method further comprising the step of using the power control channel for controlling, at the second rate, power on the reverse channels of a plurality of terminals in the second state.

Preferably the forward channels include a control channel for identifying, for each time slot of the tdm channel, a terminal to which traffic in the time slot is being communicated, and the terminals monitor said control channel at said first rate in both the first and second states. This method can include the step of switching a terminal from the second state to the first state in response to identification of said terminal on said control channel. The method can also include the step of switching between the first and second states of the terminal in dependence upon whether or not a data buffer, for traffic communication with the terminal via the tdm channel, is empty.

Another aspect of the invention provides a terminal for use in a wireless communications system, the terminal being operable in a first state to receive traffic for the terminal in respective time slots of a tdm channel of the system and to communicate pilot and feedback information at a first rate, and being operable in a second state to communicate said pilot and feedback information at a second rate which is less than the first rate, the terminal further being operable in the first and second states to monitor a control channel of the system to identify, for each time slot of the tdm channel, traffic in the time slot communicated to the terminal, and in response to such identification in the second state to switch to the first state.

Desirably the terminal is further operable in said first state to receive and respond to power control information at the first rate and in said second state to receive and respond to power control information at the second rate.

According to another aspect, this invention provides apparatus for use in a wireless communications system having forward and reverse channels for communicating traffic and control information with a plurality of terminals, the forward channels including a time division multiplexed (tdm) channel for communicating traffic in respective time slots to respective terminals, and the reverse channels including channels for communicating pilot and feedback information from the terminals, wherein the apparatus is operable to receive said pilot and feedback information from each of a plurality of terminals in a first state at a first rate, and to receive said pilot and feedback information from each of a plurality of terminals in a second state, at a second rate which is less

than the first rate, via a channel shared by said plurality of terminals in the second state.

Desirably the apparatus is further operable to supply power control information, via a shared power control channel, at the first rate to terminals in said first state and at the second rate to terminals in said second state.

A further aspect of the invention provides a method for transmitting control information from a network apparatus to a plurality of mobile stations via a common control channel, the method comprising: transmitting control information to a first set of mobile stations at a first rate of transmission via the common control channel; and transmitting control information to a second set of mobile stations at a second rate of transmission via the common control channel, the second rate of transmission being less than the first rate of transmission. The common control channel can be a common power control channel.

According to another aspect, the invention provides a wireless communications system comprising: first and second sets of mobile stations; and an apparatus that operates to transmit control information to, and/or to receive control information from, the first and second sets of mobile stations at respective first and second rates of transmission via a shared control channel, the second rate of transmission being less than the first rate of transmission. The shared control channel can comprise a power control channel for controlling power of the mobile stations.

## Brief Description of the Drawings

The invention will be further understood from the

following description by way of example with reference to the accompanying drawings, in which:

Figs. 1 and 2 represent information in communications channels of a wireless communications system in accordance with an embodiment of the invention, for a plurality of time slots and respectively for an active state and for a control hold state of an active user;

Figs. 3, 4, and 5 illustrate configurations of power control information in a time slot of a common power control channel; and

Fig. 6 illustrates a simplified state diagram for a base station and a mobile station in a control hold state and for a transition to an active state.

## Detailed Description

In an embodiment of the invention described in detail below with reference to the drawings, an active user can be in an active state or a control hold state. The active state of an active user is also referred to as a primary active state, and the control hold state of an active user is also referred to as a secondary active state. The control hold state generally corresponds to situations in which there is temporarily no packet data being communicated between the BS and the MS of an active user, i.e. in which data buffers are empty. Transitions between the active state and the control hold state can therefore be dependent upon whether or not data buffers are empty, or can be directed by the system.

An active user in this embodiment of the invention refers to a mobile station (MS), and the description below refers particularly to high rate packet data communications to

the MS on the forward link from a base station (BS), and power control for communications on the reverse link from the MS to the BS. However, it can be appreciated that these particulars are provided by way of example and not limitation, and that the invention can also be applied to other shared communications channels for traffic and/or control (including feedback, pilot, and any other overhead) information in either direction or in both directions between base stations and remote terminals.

In the described embodiment of the invention, forward and reverse link channels and system resources provided for supporting the forward link packet data communications to an active user with a packet session connected include the forward link channels F-PDCH, F-PDCCH, and F-CPCCH, the reverse link channels R-PICH, R-CQICH, and R-ACKCH, and the identifier MAC_ID, as discussed in the Background above. It is observed that the channel F-PDCH could instead be any data channel shared in a time division multiplexed manner by active users, and the channel F-PDCCH could instead be any shared control channel for conveying to the active users information required for detecting their respective data on the shared data channel. In addition, in the described embodiment of the invention the common power control channel F-CPCCH is assumed to be defined in a manner similar to that of known cdma2000 systems, with 24 bits in each time slot, so that a respective bit in successive time slots can constitute a sub-channel for controlling reverse link power of a respective MS.

Also in the described embodiment of the invention it is assumed that the time slot duration is 1.25 ms, the time slots having a rate of 800 Hz which is referred to as the full rate. Other, lower, rates referred to below are conveniently sub-multiples of the full rate, for example 1/2 (half), 1/4

(quarter), and 1/8 (eighth) rates being 400, 200, and 100 Hz respectively and corresponding to one in every two, four, and eight time slots respectively. However, it can be appreciated that these parameters are given only by way of example.

The MS of an active user in the active state operates at the full rate, continuously monitoring the F-PDCCH (shared control channel) for its identifier MAC_ID to determine when it should detect data on the F-PDCH, and monitoring its assigned sub-channel of the F-CPCCH at the full rate (one bit per time slot of the common power control channel) for power control of its transmissions on the reverse link. It also sends feedback information on the reverse link, i.e. the R-PICH, R-CQICH, and R-ACKCH channels, at the full rate of 800 Hz. These full rate communications to and from the MS are known in the art and are not further described here.

The MS of an active user in the control hold state operates in several respects at a reduced or lower rate, for example at a selected one of the 1/2, 1/4, and 1/8 rates already mentioned. More particularly, in the control hold state the MS still continuously monitors the F-PDCCH (shared control channel) for its identifier MAC_ID, but it can monitor the F-CPCCH at the reduced rate as described below. In addition, it sends its pilot and feedback information on the reverse link channels R-PICH and R-CQICH at the reduced rate, and does not send any acknowledgements on the R-ACKCH channel because in the control hold state there is no received data to be acknowledged.

It can be appreciated that the greatest advantages are provided by adopting reduced rates for all channels on the forward link and on the reverse link for which such reduced rates are possible. However, this need not be the case and

reduced rates may be adopted for only some of these channels. For example, it can be appreciated that reduced rates may be adopted for the reverse link channels without also adopting a reduced rate for the F-CPCCH on the forward link. In addition, it can be appreciated that different reduced rates may be adopted for different active users, and/or at different times for the same active user.

As a result of the reduced rates used in the control hold state, in accordance with embodiments of this invention system resources that previously could be used by only one active user can be shared by a plurality of active users each in the control hold state, so that a greater total number of active users can be simultaneously supported by the system. Transitions between the active state and the control hold state for each active user can take place relatively rapidly (compared with switching between idle and active states), as described further below.

Referring to Fig. 1, information in the shared forward link channels F-PDCH, F-PDCCH, and F-CPCCH referred to above, and in the reverse link channels R-PICH, R-CQICH, and R-ACKCH dedicated to an individual active user in the active state, is represented by hatching for a plurality of time slots numbered 1 to 7. As described above, the MS in the active state continuously monitors the F-PDCCH for its assigned MAC_ID, for example id1 in the time slot 2 in Fig. 1, and detects and decodes the data in the corresponding time slot of the high rate packet data channel F-PDCH, "data for id1" in Fig. 1. The other time slots of these shared channels as shown in Fig. 1 contain data and identities for other active users which are in the active state.

The MS in the active state also continuously monitors the shared F-CPCCH, to detect in an assigned one of (in this embodiment) 24 bit positions in each time slot the bits 10 of the power control sub-channel for this MS, and uses this information in known manner to control the power of its transmissions on the reverse link channels.

As illustrated in Fig. 1, the MS in the active state transmits its pilot and feedback information in each time slot in the dedicated reverse link channels.

Fig. 2 similarly illustrates by hatching information in the shared forward link channels F-PDCH, F-PDCCH, and F-CPCCH, and in the reverse link channels R-PICH, R-CQICH, and R-ACKCH of an individual active user in the control hold state, for a plurality of time slots also numbered 1 to 7. As described above, the MS in the control hold state continuously monitors the F-PDCCH for its assigned MAC_ID, which in this case is not present because no data for the MS is present on the F-PDCH, the information on each of these shared channels relating to other users.

For an MS in the control hold state, the shared F-CPCCH contains power control information for the MS at a reduced rate of 1/4 as shown in Fig. 2, in bits 12 of every fourth time slot. The MS in the control hold state monitors only these bits of the reduced rate power control sub-channel assigned to it as further described below, and uses this information in known manner to control the power of its transmissions on the reverse link channels. The same bits in other time slots can be used for other MSs in the control hold state, thereby increasing the number of active users for which power control can be supported. Other bits in the time slots of the F-CPCCH are used for power control of other MSs of

active users either in the active state as described above or similarly in the control hold state.

As also illustrated in Fig. 2, the MS in the control hold state does not transmit any acknowledgements on the R-ACKCH because there is no data to be acknowledged, and transmits its pilot and channel quality feedback information on the channels R-PICH and R-CQICH at the same reduced rate of 1/4, in this case in the time slots numbered 1 and 5. The same reverse link channels can be simultaneously used, in the other time slots and at the same or a different reduced rate, by the MSs of other active users in the control hold state, thereby also facilitating an increase of the number of active users that can be supported without additional system resources.

Figs. 3, 4, and 5 illustrate by way of example three possible fixed configurations of the information in each time slot of the F-CPCCH, which is divided into different fields 14 and 16 for power control for MSs in the active state and in the control hold state respectively. It can be appreciated that other fixed configurations can be used, and/or the configuration can be changed dynamically for example in dependence upon the real time traffic load of active users in the system, with updated configuration information being sent on a shared control channel such as the F-PDCCH.

In the configuration of Fig. 3, the field 14 contains 20 bits to provide power control sub-channels for 20 MSs in the active state, in known manner with one bit per user in each time slot as described above. The field 16 contains the remaining 4 bits in each time slot, each bit being used to provide power control for different MSs in the control hold state in the reduced rate manner described above. For example if the reduced rate is 1/8, then in a 10 ms cycle of 8 time

slots this field provides one bit for power control of each of 32 MSs in the control hold state, so that the F-CPCCH can support a total of 52 active users.

In the configuration of Fig. 4, the field 14 contains 16 bits to provide power control sub-channels for 16 MSs in the active state, in known manner with one bit per user in each time slot as described above. The field 16 contains the remaining 8 bits in each time slot, each bit being used to provide power control for different MSs in the control hold state in the reduced rate manner described above. For example if the reduced rate is 1/4, then in a 5 ms cycle of 4 time slots this field provides one bit for power control of each of 32 MSs in the control hold state, so that the F-CPCCH can support a total of 48 active users.

In the configuration of Fig. 5, the field 14 also contains 16 bits to provide power control sub-channels for 16 MSs in the active state, in known manner with one bit per user in each time slot. The field 16 contains the remaining 8 bits in each time slot, each bit being used to provide power control for different MSs in the control hold state in the reduced rate manner described above, using 4 bits for power control of MSs operating at each of two alternative reduced rates. For example if the two alternative reduced rates are 1/8 and 1/4, then this field provides bits for power control of each of 4(4+8)=48 MSs in the control hold state, so that the F-CPCCH can support a total of 64 active users.

Any of a number of different techniques may be used to switch an MS of an active user between the active state and the control hold state. For example, switching from the active state to the control hold state can be dependent, immediately or after a small delay, upon a data buffer at the BS for an

active user becoming empty, and switching from the control hold state to the active state can be dependent upon the data buffer ceasing to be empty and/or the data buffer occupancy being higher than a certain threshold. Such switching can be implemented using Layer 2 and/or Layer 3 messaging or signalling.

For example, in an embodiment of the invention, if a data buffer, for sending data on the forward link to an MS in the active state, becomes empty as a result of a data packet being sent from the BS to the MS, the Layer 2 frame header of this packet can include a buffer empty indication (1 bit) and information (a mode control field) regarding the control hold state. In response to receipt of the forward link buffer empty indication, the MS in the active state replies by indicating whether or not the sending data buffer at the MS is also empty. If it is empty, the active user is switched from the active state to the control hold state immediately after the response from the MS is received. Otherwise, the MS remains in the active state.

Alternatively, a timer can be set with a very small value when the data buffer for sending data to the MS in the active state becomes empty. When the set time ends, or at any other time that the network considers appropriate, the network can for example send a control message in the Layer 2 frame header to explicitly request the MS to switch to the control hold mode. Alternatively, Layer 3 signaling can be used to trigger a switch of an active user from the active state to the control hold state.

By way of example of a mode control field in the Layer 2 frame header as mentioned above, such a mode control field can comprise a time slot indicator and a sub-channel

indicator, and optionally a reduced rate indicator. For example, for the configuration described above with reference to Fig. 3 with a reduced rate of 1/8, the time slot indicator may comprise 3 bits to indicate which time slot, in a sequence of 8 time slots in the 10 ms cycle, is to be used by the MS in the control hold state, and the sub-channel indicator may comprise 2 bits to indicate which of the 4 bits of the field 16 is to be used by the MS for its reduced rate power control sub-channel.

Similarly, for the configuration described above with reference to Fig. 4 with a reduced rate of 1/4, the time slot indicator may comprise 2 bits to indicate which time slot, in a sequence of 4 time slots in the 5 ms cycle, is to be used by the MS in the control hold state, and the sub-channel indicator may comprise 3 bits to indicate which of the 8 bits of the field 16 is to be used by the MS for its reduced rate power control sub-channel.

Also, for the configuration described above with reference to Fig. 5 with alternative reduced rates of 1/8 and 1/4, there may be a 1-bit reduced rate indicator to indicate to the MS which of the two reduced rates is to be used, the time slot indicator may comprise 2 or 3 bits to indicate which time slot, in a sequence of either 4 or 8 time slots for the 1/4 and 1/8 reduced rates respectively, is to be used by the MS, and the sub-channel indicator may comprise 2 bits in each case to indicate which of the respective 4 bits of the field 16 is to be used by the MS for its reduced rate power control sub-channel.

Thus it can be appreciated that the configuration information on the F-PDCCH and the mode control field on the

F-PDCH together provide the MS of an active user with rules for its operation in the control hold state.

It can be appreciated that this information can be reduced if the MS operates at a reduced rate in the control hold state for only some of the possible reduced rate channels discussed above. For example, in the event that the common power control channel F-CPCCH is monitored continuously in the control hold state of an MS, as described above with reference to Fig. 1, even though the reduced rate monitoring of this channel could be adopted as described above with reference to Fig. 2, the mode control field discussed above need not necessarily contain an F-CPCCH sub-channel indicator (for example, the power control sub-channel previously assigned to the MS in the active state may continue to be used in the control hold state). In this case, in the same manner as described above, the time slot indicator and optional reduced rate indicator can inform the MS which time slots to use for the reduced rate information on the R-PICH and R-CQICH as described above with reference to Fig. 2.

A BS can send data to an MS in the control hold state at any time, causing the MS to switch to the active state, because the MS continuously monitors the shared control channel F-PDCCH. Thus in response to the sending data buffer, for an active user whose MS is in the control hold state, becoming non-empty, the BS can send data to the MS immediately on the F-PDCH, using parameters (such as data rate and modulation scheme) determined from the most recent R-CQICH information from the MS. A control field of for example 5 bits can be inserted into the header of the Layer 2 frame which encapsulates the first data packet, in order to inform the MS which sub-channel (i.e. which bit in each time slot) of the

F-CPCCH it should monitor once it is in the active state, and the BS starts to send this power control sub-channel at the full rate at the same time. The MS detects its MAC_ID in the F-PDCCH which it has continued to monitor in the control hold state, and in response switches to the active state and detects and decodes the packet data intended for it on the F-PDCH. Accordingly, the MS now in the active state sends its R-PICH, R-CQICH, and R-ACKCH at the full rate as described above and shown in Fig. 1. In conventional manner, the BS re-sends packet data for which it receives a NAK or no acknowledgement within a certain time period.

Alternatively, an MS in the control hold state can be switched to the active state by simultaneous transmission to it of a data packet and Layer 3 signaling sent in an assured mode, the MS accordingly returning to full rate transmission of its channels R-PICH, R-CQICH, and R-ACKCH.

Fig. 6 illustrates a state diagram for a BS and an MS, illustrating operations in the control hold state, the active state, and for a transition from the control hold state to the active state initiated by data for the MS arriving at the BS, or by Layer 3 signaling. It can be appreciated that this provides a simplified illustration by way of example of the respective states and transitions, and that transitions from the active state to the control hold state, and transitions requested by the MS, can be similarly illustrated.

As shown in Fig. 6, and as described above, in the control hold state for the MS of an active user the BS receives the reduced-rate (also referred to as gated) reverse link channels R-PICH and R-CQICH which are sent by the MS. Also as described above, in this state the BS advantageously sends reduced-rate power control commands on the F-CPCCH, and these

are received by the MS. The MS also continuously (i.e. at the full rate) monitors the F-PDCCH for its assigned MAC_ID, to determine if and when there is data for the MS on the F-PDCH.

In response to data for the MS arriving at the BS (the sending data buffer for the MS becomes non-empty), or Layer 3 signaling, the BS proceeds to a transition phase from the control hold state to the active state for the MS. As shown in Fig. 6, in this transition phase the BS sends the data or signaling on the F-PDCH (and correspondingly sends the MAC_ID for the MS on the F-PDCCH as described above), starts a transition timer, and monitors the channel R-ACKCH for an acknowledgement of the data packet sent, with possible retransmission in the event that no ACK is received within a timeout period. In the event that the transition timer expires, the BS proceeds to a call clearing procedure. Also in the transition phase the BS monitors the duty cycle of the R-PICH and R-CQICH to detect a transition from the reduced-rate (gated) transmission to full rate transmission of these channels by the MS.

Correspondingly, the MS detects its MAC_ID on the F-PDCCH and receives the corresponding data or signaling on the F-PDCH. Consequently, in the transition phase the MS resumes sending the R-PICH, R-CQICH, and R-ACKCH at the full rate, and sends an acknowledgement on the R-ACKCH.

On detection of the full-rate R-PICH and R-CQICH, or, in the case of Layer 3 signaling, at an action time determined by this signaling, the BS resumes full-rate transmission of power control commands for the MS via the F-CPCCH in a new time slot or power control sub-channel, and the MS uses this full-rate power control accordingly. In addition, the BS resumes

continuous or full-rate monitoring of the reverse link channels R-PICH, R-CQICH, and R-ACKCH.

Although embodiments of the invention are described above in the context of a particular type of wireless communications system and for particular channels such as a power control channel, pilot channel, and channel quality feedback channel, it should be understood that the invention is not limited to these and can be applied to any of these channels and/or other channels, in this or and/or other types of system.

Thus although particular embodiments of the invention and variations have been described above in detail, it can be appreciated that numerous modifications, variations, and adaptations may be made within the scope of the invention as defined in the claims.

CLAIMS:

1. A method of communicating control information in a wireless communications system, comprising the steps of:

in a first state of a terminal for traffic communication with the terminal, communicating control information with the terminal at a first rate;

in a second state of the terminal, communicating at least some of said control information at a second rate which is less than the first rate; and

sharing a communication channel for the control information at the second rate for communicating control information among a plurality of terminals in the second state.

2. A method as claimed in claim 1 wherein the control information communicated at the second rate in the second state of the terminal comprises power control information for the terminal.

3. A method as claimed in claim 1 or 2 wherein the control information communicated at the second rate in the second state of the terminal comprises a pilot and/or a channel quality indication from the terminal.

4. A method as claimed in claim 1, 2, or 3 wherein the control information is communicated in time slots of at least one communication channel of the system, wherein control information at the first rate is communicated with a terminal in the first state in each time slot on the communication channel, and wherein control information at the second rate is communicated with a terminal in the second state in only one of every N time slots on the communication channel, where N is an integer greater than one.

5. A method of communicating control information in a wireless communications system having forward and reverse channels for communicating traffic and control information with a plurality of terminals, the forward channels including a time division multiplexed (tdm) channel for communicating traffic in respective time slots to respective terminals, and the reverse channels including channels for communicating pilot and feedback information from the terminals, the method comprising the steps of:

in a first state of a terminal for receiving traffic for the terminal in respective time slots of the tdm channel, communicating said pilot and feedback information at a first rate;

in a second state of the terminal, communicating said pilot and feedback information at a second rate which is less than the first rate; and

sharing the channels for communicating said pilot and feedback information at the second rate among a plurality of terminals in the second state.

6. A method as claimed in claim 5 wherein the first rate is equal to a rate of said time slots and the second rate is 1/2, 1/4, or 1/8 of the first rate.

7. A method as claimed in claim 5 or 6 wherein the forward channels include a power control channel for controlling, at the first rate, power on the reverse channels of a respective terminal in the first state, the method further comprising the step of using the power control channel for controlling, at the second rate, power on the reverse channels of a plurality of terminals in the second state.

8. A method as claimed in claim 5, 6, or 7 wherein the forward channels include a control channel for identifying, for each time slot of the tdm channel, a terminal to which traffic in the time slot is being communicated, and the terminals monitor said control channel at said first rate in both the first and second states.

9. A method as claimed in claim 8 and including the step of switching a terminal from the second state to the first state in response to identification of said terminal on said control channel.

10. A method as claimed in any of claims 1 to 9 and including the step of switching between the first and second states of the terminal in dependence upon whether or not a data buffer for traffic communication with the terminal is empty.

11. A terminal for use in a wireless communications system, the terminal being operable in a first state to receive traffic for the terminal in respective time slots of a tdm channel of the system and to communicate pilot and feedback information at a first rate, and being operable in a second state to communicate said pilot and feedback information at a second rate which is less than the first rate, the terminal further being operable in the first and second states to monitor a control channel of the system to identify, for each time slot of the tdm channel, traffic in the time slot communicated to the terminal, and in response to such identification in the second state to switch to the first state.

12. A terminal as claimed in claim 11, the terminal being further operable in said first state to receive and respond to power control information at the first rate and in said second

state to receive and respond to power control information at the second rate.

13. Apparatus for use in a wireless communications system having forward and reverse channels for communicating traffic and control information with a plurality of terminals, the forward channels including a time division multiplexed (tdm) channel for communicating traffic in respective time slots to respective terminals, and the reverse channels including channels for communicating pilot and feedback information from the terminals, wherein the apparatus is operable to receive said pilot and feedback information from each of a plurality of terminals in a first state at a first rate, and to receive said pilot and feedback information from each of a plurality of terminals in a second state, at a second rate which is less than the first rate, via a channel shared by said plurality of terminals in the second state.

14. Apparatus as claimed in claim 13 and being further operable to supply power control information, via a shared power control channel, at the first rate to terminals in said first state and at the second rate to terminals in said second state.

15. A method for transmitting control information from a network apparatus to a plurality of mobile stations via a common control channel, the method comprising:

transmitting control information to a first set of mobile stations at a first rate of transmission via the common control channel; and

transmitting control information to a second set of mobile stations at a second rate of transmission via the common

control channel, the second rate of transmission being less than the first rate of transmission.

16. A method according to claim 15, wherein the common control channel is a common power control channel.

17. A wireless communications system comprising:

first and second sets of mobile stations; and

an apparatus that operates to transmit control information to, and/or to receive control information from, the first and second sets of mobile stations at respective first and second rates of transmission via a shared control channel, the second rate of transmission being less than the first rate of transmission.

18. A wireless communications system as claimed in claim 17 wherein the shared control channel comprises a power control channel for controlling power of the mobile stations.

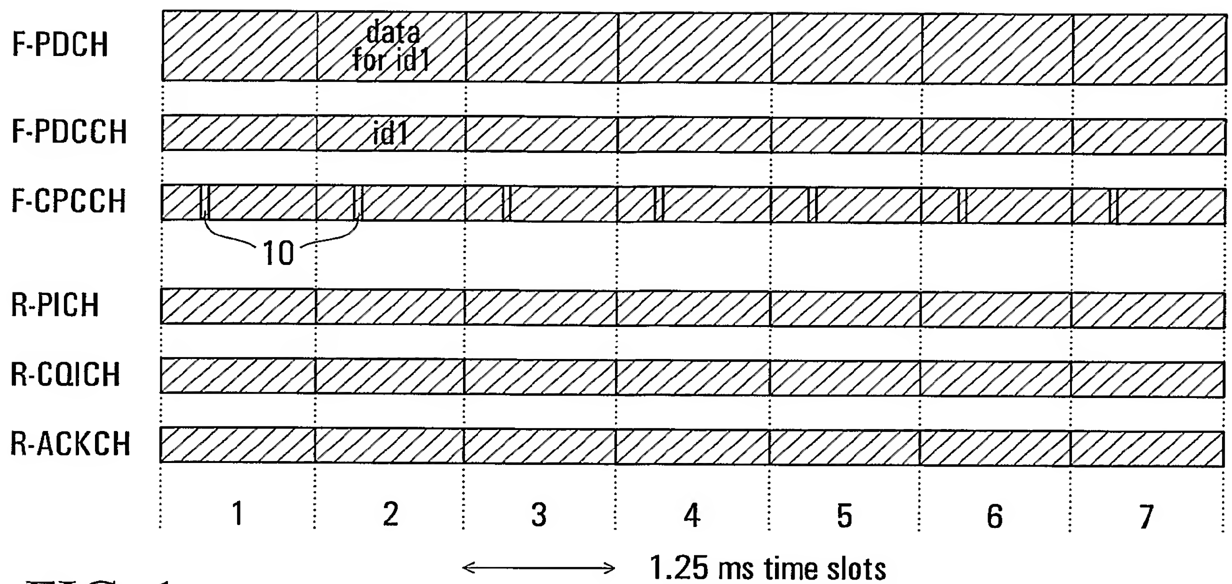

F-PDCH
data for id1
F-PDCCH
id1
F-CPCCH
10
R-PICH
R-CQICH
R-ACKCH
1
2
3
4
5
6
7
1.25 ms time slots

FIG. 1

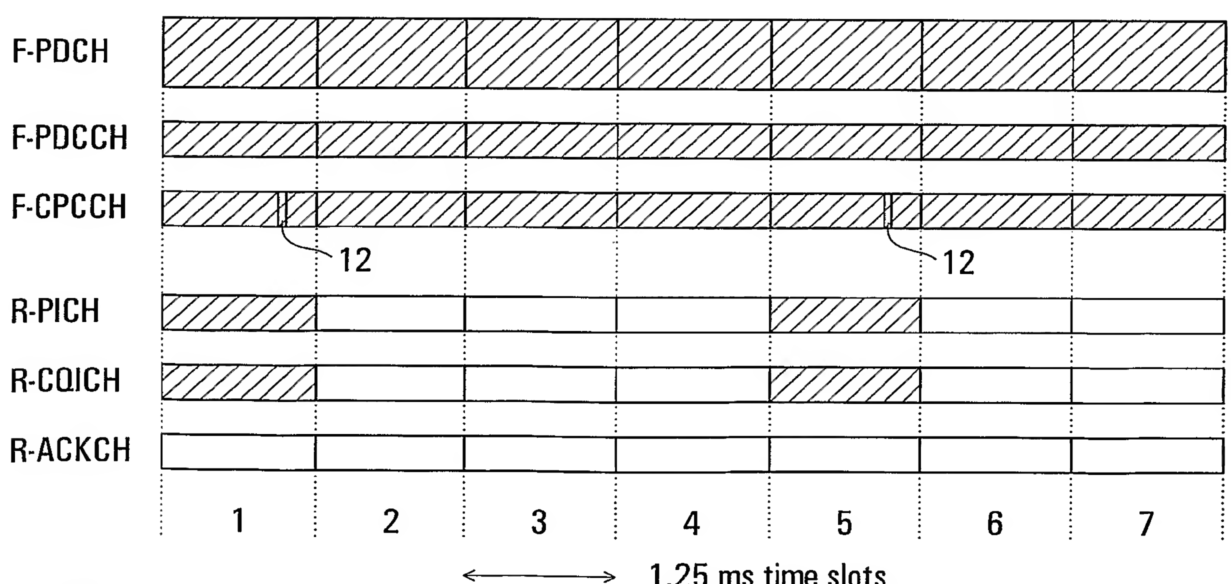

F-PDCH
F-PDCCH
F-CPCCH
12
12
R-PICH
R-CQICH
R-ACKCH
1
2
3
4
5
6
7
1.25 ms time slots

FIG. 2

FIG. 3

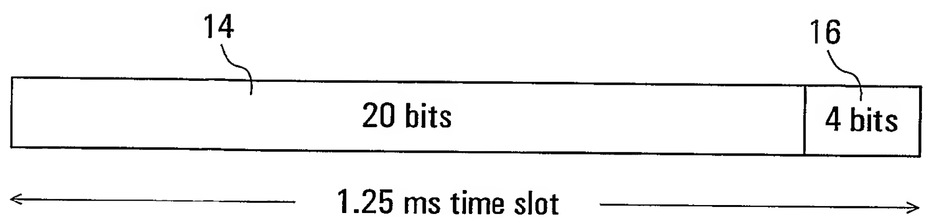



FIG. 4

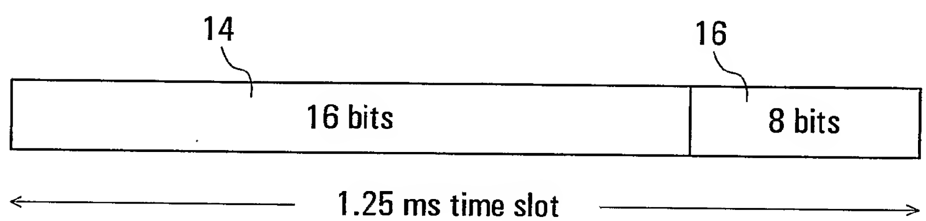
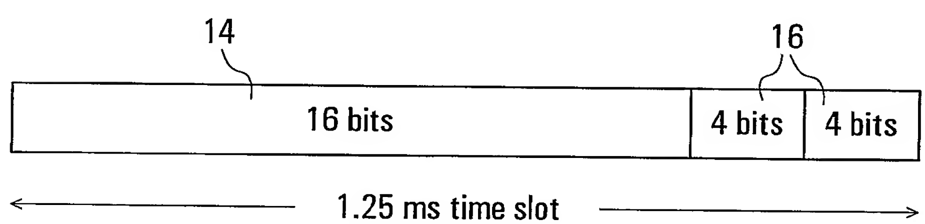



FIG. 5



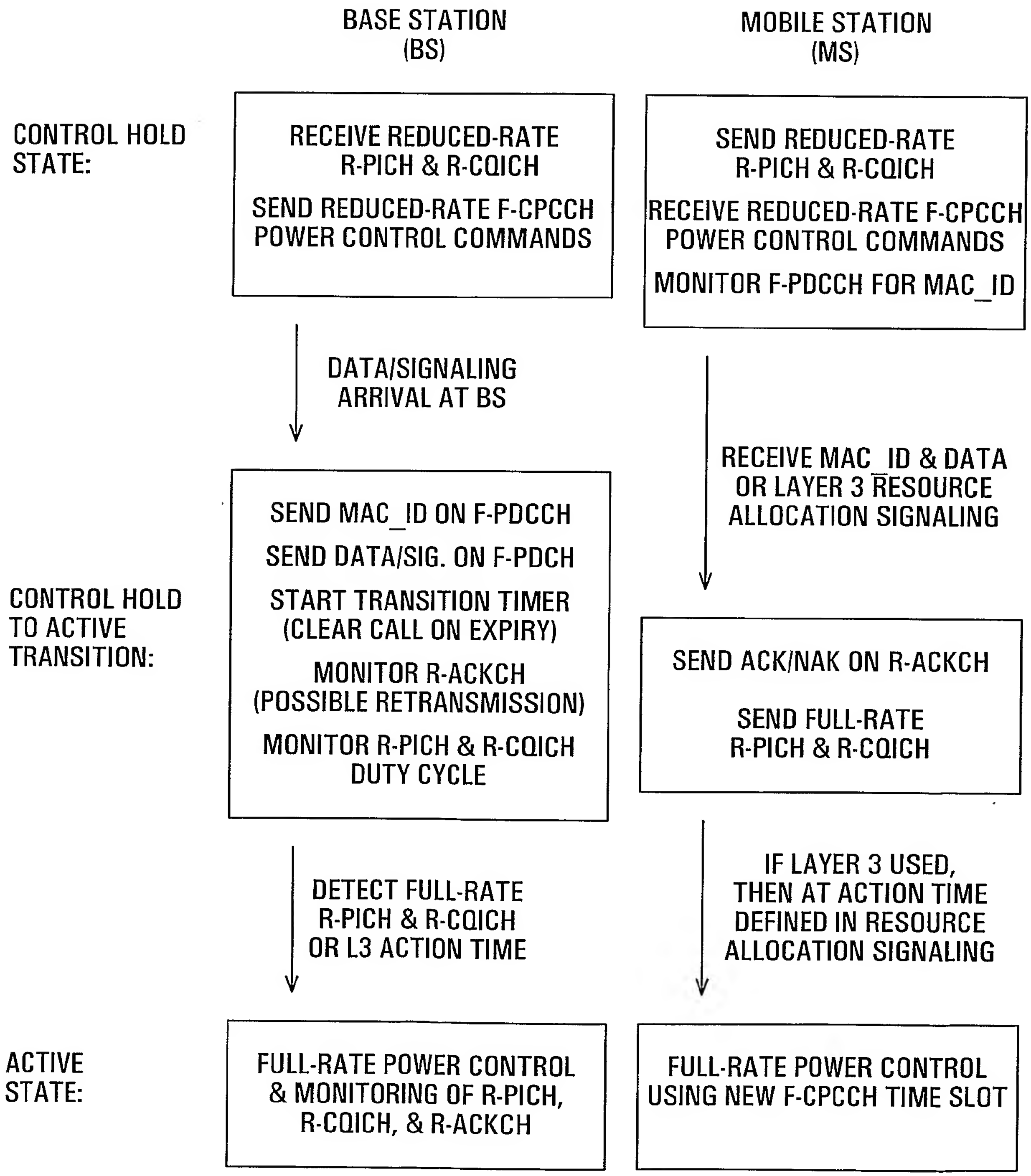

BASE STATION
(BS)
MOBILE STATION
(MS)
CONTROL HOLD
STATE:
RECEIVE REDUCED-RATE
R-PICH & R-CQICH
SEND REDUCED-RATE F-CPCCH
POWER CONTROL COMMANDS
SEND REDUCED-RATE
R-PICH & R-CQICH
RECEIVE REDUCED-RATE F-CPCCH
POWER CONTROL COMMANDS
MONITOR F-PDCCH FOR MAC_ID
DATA/SIGNALING
ARRIVAL AT BS
RECEIVE MAC_ID & DATA
OR LAYER 3 RESOURCE
ALLOCATION SIGNALING
CONTROL HOLD
TO ACTIVE
TRANSITION:
SEND MAC_ID ON F-PDCCH
SEND DATA/SIG. ON F-PDCH
START TRANSITION TIMER
(CLEAR CALL ON EXPIRY)
MONITOR R-ACKCH
(POSSIBLE RETRANSMISSION)
MONITOR R-PICH & R-CQICH
DUTY CYCLE
SEND ACK/NAK ON R-ACKCH
SEND FULL-RATE
R-PICH & R-CQICH
DETECT FULL-RATE
R-PICH & R-CQICH
OR L3 ACTION TIME
IF LAYER 3 USED,
THEN AT ACTION TIME
DEFINED IN RESOURCE
ALLOCATION SIGNALING
ACTIVE
STATE:
FULL-RATE POWER CONTROL
& MONITORING OF R-PICH,
R-CQICH, & R-ACKCH
FULL-RATE POWER CONTROL
USING NEW F-CPCCH TIME SLOT

FIG. 6

# INTERNATIONAL SEARCH REPORT

In onal Application No
PCT/CA 02/00941

**A. CLASSIFICATION OF SUBJECT MATTER**
IPC 7 H04Q7/22 H04L12/28 H04L12/56 H04B7/26 H04Q7/38

According to International Patent Classification (IPC) or to both national classification and IPC

**B. FIELDS SEARCHED**

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04Q H04L H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal

**C. DOCUMENTS CONSIDERED TO BE RELEVANT**

| Category ° | Citation of document, with indication, where appropriate, of the relevant passages | Relevant to claim No. |
|---|---|---|
| X | WO 00 62456 A (SAMSUNG ELECTRONICS CO LTD) 19 October 2000 (2000-10-19)<br>abstract<br>page 1, line 14 - line 28<br>page 12, line 14 -page 32, line 35<br>figures 1A-12E<br>--- | 1-18 |
| X | WO 00 35126 A (SAMSUNG ELECTRONICS CO LTD) 15 June 2000 (2000-06-15)<br>page 1, line 9 -page 6, line 17<br>page 9, line 15 -page 44, line 5<br>figures 1-6<br>----- | 1-18 |

☐ Further documents are listed in the continuation of box C.

☒ Patent family members are listed in annex.

° Special categories of cited documents :

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"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

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"&" document member of the same patent family

| Date of the actual completion of the international search | Date of mailing of the international search report |
|---|---|
| 11 October 2002 | 24/10/2002 |

| Name and mailing address of the ISA | Authorized officer |
|---|---|
| European Patent Office, P.B. 5818 Patentlaan 2<br>NL – 2280 HV Rijswijk<br>Tel. (+31–70) 340–2040, Tx. 31 651 epo nl,<br>Fax: (+31–70) 340–3016 | Martinozzi, A |

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| | | CN 1347604 T | 01-05-2002 |
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| | | WO 0035126 A1 | 15-06-2000 |